OTS: 60-41,184

JPRS: 5318

14 November 1960

EXCERPTS FROM "COMMUNICATIONS CABLE"

By I. I. Grodnev
and
B. F. Miller

- USSR -

Distributed by:

OFFICE OF TECHNICAL SERVICES
U. S. DEPARTMENT OF COMMERCE
WASHINGTON 25, D. C.

U. S. JOINT PUBLICATIONS RESEARCH SERVICE
1636 CONNECTICUT AVE., N. W.
WASHINGTON 25, D. C.

F O R E W O R D

This publication was prepared under contract by the UNITED STATES JOINT PUBLICATIONS RESEARCH SERVICE, a federal government organization established to service the translation and research needs of the various government departments.

JPRS: 5318

CSO: 4528-N

EXCERPTS FROM "COMMUNICATIONS CABLE"

/Following is the translation of excerpts by I. I. Grodnev and B. F. Miller from "Kabeli Svyazi" (English version above), Moscow, 1950, pages 61-350/

CHAPTER THREE

SYMMETRICAL CABLES

A. Cable Designs and Their Calculation

3-1. Elements of the Designs

The conductors of symmetrical cables usually consist of round annealed copper wire isolated by a concentric layer of insulation.

The diameters of the wires in use are presented in Table 1.

A composition of paper and air which possesses a relatively low dielectric constant and thereby provides for sufficiently low attenuation per kilometer is used as insulation for the conductors of symmetrical main cables which are employed to transmit audio and higher frequencies over considerable distances.

In short cables that branch off from the mains (e. g., in telephone distribution cables and office cables), and in holding and signalling cables used for transmission of d-c pulses, use is made of solid impregnated fiber insulation (made from paper, cotton, or silk fibers), frequently in combination with a thin layer of enamel which is applied to the wire. Impregnation lessens the hygroscopic tendencies of the insulation and raises its dielectric strength.

The latter is of particular importance for holding cables. The application of the air-and-paper insulation to a cable conductor may be accomplished by any of four different methods.

a) With a longitudinal strip which forms a trihedral prism (triangle) about the wire and is wound with a loose spiral of cotton thread for strength. b) With a closed paper spiral wound about the wire to form one or two hollow tubes, which are compressed until they wrinkle. c)

By covering the wire with a continuous porous paper mass ("paper-pulp" insulation). d) By winding the wire with an open spiral of packthread (paper thread) with one or two paper strips applied over it. The technology of insulating by the methods indicated is discussed in Chapter 10.

All types of air-and-paper insulation are represented schematically in Fig. 3-1. Table 3-1 also indicates the ranges of their application.

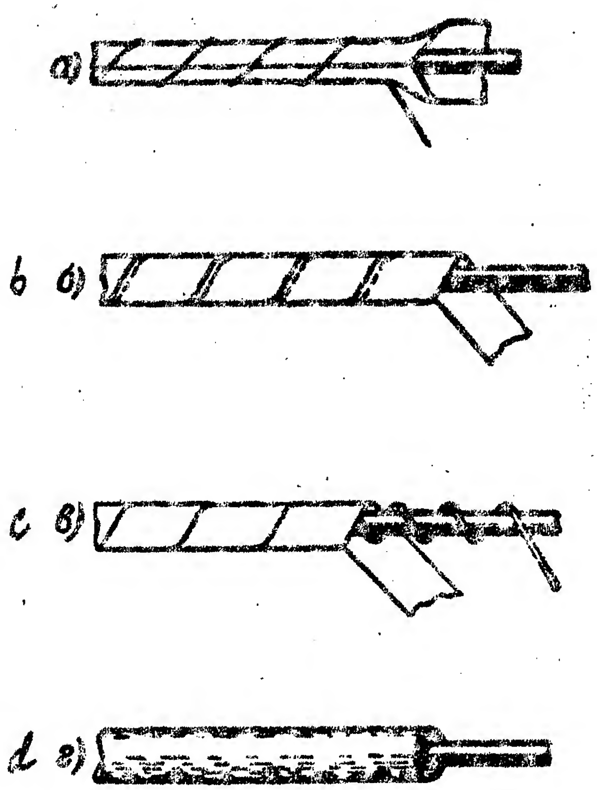


Fig. 3-1. Types of insulation

a) air-and-paper insulation, trihedral; b) air-

and-paper insulation in hollow winding (tube); c) paper-pulp insulation; d) packthread-paper insulation.

The individual conductors are usually combined (twisted) into groups which are called the elements of the symmetrical cable. Twisting places the individual conductors of the working circuit under similar conditions with respect to all external interference and facilitates their displacement with respect to one another when the cable is flexed. The following types of elements are used in existing communications cables:

a) the pair (P);

b) the spiral quad (Z);

c) the double pair (DP);

d) the sextuplet;

e) the octuplet or double quad (DZ);

f) the triplet.

The first three elements have found the most widespread use. The others are used much more rarely and, for the most part, due to design considerations. In certain cases, the individual elements are wound with an additional annular layer of insulation or screening tape. The former are referred to as reinforced and the latter as screened elements. The special system of conventional designations

SYMMETRICAL CABLE TYPES Table 3-1

Type of Communications Cable	Wire diam., mm	Types of insulation	System in which elements are twisted
Urban telephone cable (mains)	0.5, 0.6, and 0.7	1. Longitudinal paper-air 2. Hollow paper winding 3. Paper-pulp	Paired Spiral quad
Telephone distribution cable	0.5	1. Enamel + one winding of impregnated cotton thread 2. Two layers of impregnated cotton thread	Paired
Telephone office cable	0.5	1. Enamel + one winding of cotton thread 2. Two layers of impregnated cotton thread 3. Layer of impregnated silk and impregnated cotton thread	Paired Tripled Quadded
Long-distance communications cables (packthread type)	0.8; 0.9; 1.0; 1.2; and 1.4	Packthread-paper	Paired, spiral quad and double pair
Signaling and holding cables	1.0	Solid impregnated winding by paper strip	Single conductors

presented in Table 3-2 has been worked out for convenience in schematic representation of the cross-sections of the various cables and to simplify their classification.

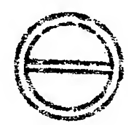
The final twisting operation, i.e., the joining of the elements to form the cable, is either carried out by lays (concentric layers) or else indicidual groups (strands or bunches) are first twisted from individual elements and then the entire cable spun from these groups (bunched twisting).

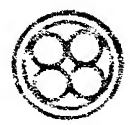
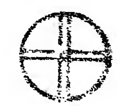
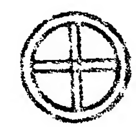
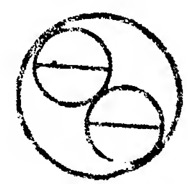
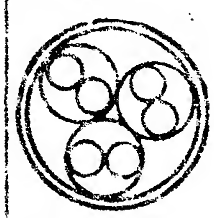
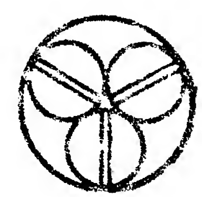
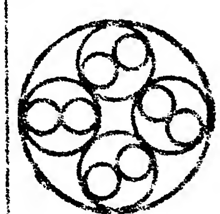
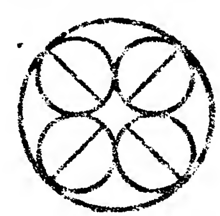
In bunched twisting, the separate bunches are usually twisted by the "random" method, i.e., all elements are twisted simultaneously in the same direction and with the same step (lay).

Key to Table 3-2:

1. CONVENTIONAL DESIGNATIONS AND GRAPHIC REPRESENTATIONS OF INDIVIDUAL COMMUNICATIONS-CABLE ELEMENTS
2. Schematic section
3. Name applied to element
4. Conventional designation (D is the diameter of the wire)
5. Conventional graphic representation
6. Structural characteristics
7. Pair (P)
8. Two differently colored conductors are twisted and wound (or not wound) with an open cotton-thread spiral
9. Reinforced pair (PU)

1 **Условные обозначения и графические изображения отдельных элементов кабел**

2 ...ический разрез	3 Наименование элемента	Условное обозначение (D—диаметр проволоки) 4	Условное графическое изображение 5	Конструктивная хар... 6
	Пара (П) 7	1×2×D		Две жилы разной расцветк... (или не обмотаны) открытой с... мажной пряжи 8
	Пара усиленная (ПУ) 9	1×2у×D 9а		10 Две жилы разной расцветк... двумя слоями бумажной ленты, ...ний) может быть наложен в ... с воздушным зазором
	11 Пара экранированная (ПЭ)	12 1×2экр×D		13 Две жилы разной расцветк... двумя слоями бумажной ленты... рующей ленты. Первая (нижняя... быть заменена продольно нало... в виде трубки с воздушным з...
	14 Четверка звездная (З)	1×4×D		15 Четыре жилы разной рас... собой и обмотаны открытой ... мажной пряжи. Рабочие пары ... рально противоположных жил

	Четверка звездная усиленная (ЗУ) 16	1×4у×D 17		18 Четыре жилы разной ... обмотаны (или не обмота... хлопчатобумажной пряжи, двумя слоями бумажной лент... ной) может быть наложен с воздушным зазором
	Четверка звездная экранированная (ЗЭ) 19	1×4э×D 20		21 Четыре жилы разной рас... бой и обмотаны (или не об... из хлопчатобумажной пряжи, двумя слоями бумажной лен... рующей ленты. Первая (ниж... быть заменена продольно на... в виде трубки с воздушным
	Двойная пара (ДП) 22	1×2×2×D		Две пары, скрученные из... чены между собой в обратн... крытой спиралью из хлопчат... 23
	Шестерка 24	1×3×2×D		Три пары, скрученные и обмотанные (каждая) откр... бумажной пряжи (разного ... скручены между собой и ... мажной ленты с перекрытие... 25
	Двойная звезда (ДЗ) 26	1×4×2×D		Четыре пары, скрученные ... скручены между собой и о... из хлопчатобумажной пряжи 27
	Тройка 28	1×3×D		29 Три жилы разной расцве... и обмотаны открытой спира... пряжи

9a. 1 X 2uxD [u=reinforced]

10. Two differently colored conductors are twisted and wound with two layers of paper stripping; here, one layer (the lower) may be applied in the form of a longitudinal tube with an air gap

11. Screened pair (PE)

12. 1x2ekrxD [ekr=screened]

13. Two differently-colored conductors are twisted and wound with two layers of paper tape and one layer of screening tape. The first (inner) paper winding may be replaced by a longitudinally-applied paper strip forming a tube with an air gap

14. Spiral quad (Z)

15. Four differently colored conductors are twisted together and wound with an open cotton-thread spiral. The working pairs are formed by diametrically-opposed conductors (across a diagonal)

16. Reinforced spiral quad (ZU)

17. 1x4u x D [u=reinforced]

18. Four differently colored conductors are twisted together and wound (or not wound) with an open cotton-thread spiral. The quad is covered on the outside with two layers of paper stripping; of these, one (the inner) may be applied longitudinally in the form of a tube with an airgap

19. Screened spiral quad (ZE)

20. 1x4e x D [e=screened]

21. Four differently colored conductors are twisted together and wound (or not wound) with an open cotton-thread spiral. The quad is wound on the outside with two layers of paper stripping and one layer of screening tape. The first (inner) paper winding may be replaced

by a longitudinally-applied paper strip in the form of a tube with an airgap

22. Double pair (DP)

23. Two pairs twisted from differently colored conductors are twisted together in opposite directions and wound with an open cotton-thread spiral

24. Sextuplet

25. Three pairs twisted from differently colored conductors and (each) wound with an open spiral of cotton thread (in a different color for each pair) are twisted together and wound with two layers of paper tape, with overlap

26. Double quad (DZ)

27. Four pairs twisted from conductors of different colors are twisted together and wound with an open cotton-thread spiral

28. Triplet

29. Three differently colored conductors are twisted together and wound with an open coton-thread spiral

In concentric twisting, however, contiguous layers should have different directions (alternating directions). In layered twisting of identical elements, the number of elements in each successive layer should be larger by 6 than that in the preceding layer--i.e., if there are 2 elements in the first layer, there will be 8 in the second 14 in the third, 20 in the fourth, etc. We obtain a certain number of elements in the layers, and therefore in the cable, depending on the number of elements in the center. Table 3-3 presents possible combinations.

Table 3-3

(1) Количество элементов (или жил) по повивам и всего в кабеле, при концентрической скрутке

(2) Их количество в центре	(3) Количества по повивам (верхние цифры) Общее количество в кабеле (нижние цифры)							
	(4) Номер повива, считая от центра							
	I	II	III	IV	V	VI	VII	VIII
1	6 7	12 19	18 37	24 61	30 91	36 127	42 169	48 217
2	8 10	14 24	20 44	26 70	32 102	38 140	44 184	50 234
3	9 12	15 27	21 48	27 75	33 108	39 147	45 192	51 243
4	10 14	16 30	22 52	28 80	34 114	40 154	46 200	52 252
5	11 16	17 33	23 56	29 85	35 120	41 161	47 208	53 261

1. NUMBER OF ELEMENTS (OR CONDUCTORS) IN LAYERS AND TOTAL NUMBER IN CABLE WITH CONCENTRIC TWISTING
2. Number in center
3. Number in each layer (upper figures); total number in cable (lower figures)
4. Number of layer, reckoned from center

When communications cables are twisted from easily-deformed elements, deviations from the figures indicated in Table 3-3 amounting to 1-2 elements in either direction are permitted. This may be done with particularly little harm in urban cables with air-and-paper insulation, since deformation of these elements does not alter the parameters that have been established for them.

In packthread cables, in which the maximal capacitive-unbalance values are stipulated in advance, such departures are undesirable. In any event, departures greater than one element per lay are not admissible for these cables.

To determine the number of elements n in each layer of a composite cable, it is necessary to calculate the diameter of the central circle--the D_{ts} ($D_{central}$) of the layer in question. Then the number n of elements in the layer will be found from the equation

$$D_{ts} = \frac{1}{\pi}\sum_{1}^{\Sigma n_i} n_i d_{eff\,i}$$

where D_{ts} is the diameter of a circle drawn through the center of the elements of the layer in question;

n_i is the number of elements of the same diameter;

l is the number of other types of elements in the layer with different diameters;

$d_{eff\,i}$ is the effective diameter of an element of one of the types among the number l.

The concept of the "effective diameter d_{eff}" is characteristic for cables with loose elements twisted with a relatively long lay. When such elements are twisted, the insulation is compressed and they merge with one another to a certain extent. The latter occurs, of course, only in twisting elements having helical shapes, i.e., those

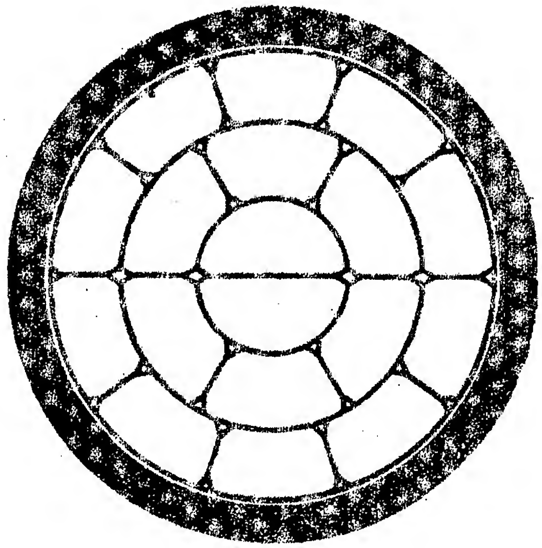


Fig. 3-2. Schematic section through cable with bunched twisting.

lacking annular windings (of the types "u" and "e"). It has been established by experiment that the d_{eff} of these elements is smaller than the diameter of the circle drawn about them. Thus, for example, with an individual conductor diameter δ,

d_{eff} for the pair = 1.65δ with a described-circle diameter of 2δ;

d_{eff} for the quad group = 2.2δ with a described-circle diameter of 2.41δ;

d_{eff} for the double-pair quad DP = 2.72δ with a described-circle diameter = 4δ;

d_{eff} for the double quad DZ = 3.63δ with a described-circle diameter of 4.84δ.

In bunched twisting of cables, the shape of the individual bundles is so deformed (Fig. 3-2) that the cable

diameters and the number of bundles in a layer can be calculated only on the basis of the areas that they occupy in the transverse section of the cable, or graphically.

The twist is characterized by its lay. The elements or conductors are arranged along helical lines. The length in which the conductor or element being twisted makes a complete revolution about the axis of the cable or element is the lay of the twist.

The ratio of the length of the twisted conductor (element) within a single lay to the total length of the lay---which indicates the factor by which the conductors (elements) in the cable are longer than the cable--is called the tightness factor or simply the "tightness."

If we unroll the surface of the twisted cable onto a plane, we obtain a right triangle (Fig. 3 3) in which one arm is the length of the cable's circumference, D, the other arm is equal in length to the lay h, and the hypotenuse is equal to the length l of the twisted conductor in one step. It follows from this triangle of evolution that the tightness

$$p = \frac{l}{h} = \frac{\sqrt{\pi^3 D^2 + h^2}}{h} = \sqrt{\pi^2 \left(\frac{D}{h}\right)^2 + 1},$$

i.e., that the tightness is a function of the ratio $\frac{h}{D}$, which is also a characteristic value for the twist. This

ratio should lie between 20 and 40 for communications cables.

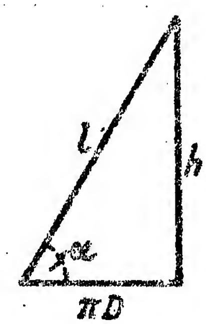


Fig. 3-3. Determination of tightness factor.

It is obviously desirable for design considerations that the ratio $\frac{h}{D}$ be large, since then the tightness factor will be smaller, and so will the outlay in all the materials composing the conductor, although this is restricted by the flexibility required of the cable.

In addition to the methods indicated above for projecting communications cables, there exists a whole series of rules, working formulas, and empirical data which have been established in the cable industry for computation of the weight of the materials and the dimensions of the separate elements of a cable design. These are considered at the end of the present chapter after a general survey of the basic types of symmetrical communications cables.

3-2. GENERAL SURVEY OF BASIC TYPES OF SYMMETRICAL COMMUNICATIONS CABLES

a) Urban Telephone-Cable Mains.

Urban telephone cables (GOST-V-1176-41) differ in number of pairs, diameter of conducting wires, and the structure of the protective sheathing. (see Table 3-4).

As will be seen from the Table, these cables are made from conductor wire 0.5, 0.6, and 0.7 mm in diameter and insulated by one of the three methods described above; Trihedral-prism, hollow-tube, or paper-pulp insulation.

The paper-pulp method of insulating the conductors is the most attractive.

In these cables, preference is given paired and, less frequently, to quadded (star) laying of the conductors.

The final twist is usually carried out by concentric layers with a definite number of pairs in a single cable (from 5 to 1200, inclusive). The standard numbers of pairs in the cables and their distribution among the layers are given in Table 3-5.

Table 3-4

(1) Городские телефонные кабели

(2) Марка кабеля	(3) Диаметр проволоки, мм		
	0,5	0,6	0,7
	(4) Число пар		
(5) ТГ	5, 10, 20, 30, 40, 50, 70, 80, 100, 150, 200, 300, 400, 500, 600, 700, 800, 900, 1 000, 1 200	(7) От 5 (8) до 1 000	От (7) 5 до 600 (8)
(6) ТА, ТБ, ТБГ и ТП	7 От 5 до 600 8	7 От 5 до 600 8	7 От 5 до 600 8
ТК	7 От 20 до 600 8	7 От 20 до 600 8	7 От 20 до 600 8

1. Urban telephone cables
2. Type of cable
3. Diameter of wire, mm
4. Number of pairs
5. TG
6. TA, TB, TBG and TP
   TK
7. from
8. to

All types of urban telephone cable have lead sheaths, the thicknesses of which are listed in Table 3-18.

Over the lead sheath the cables may have external protective coverings and armor.

Fig. 3-4 shows a 100-pair armored telephone cable of Type TB in section. The basic parameters of urban-type telephone cables are given in Table 3-6.

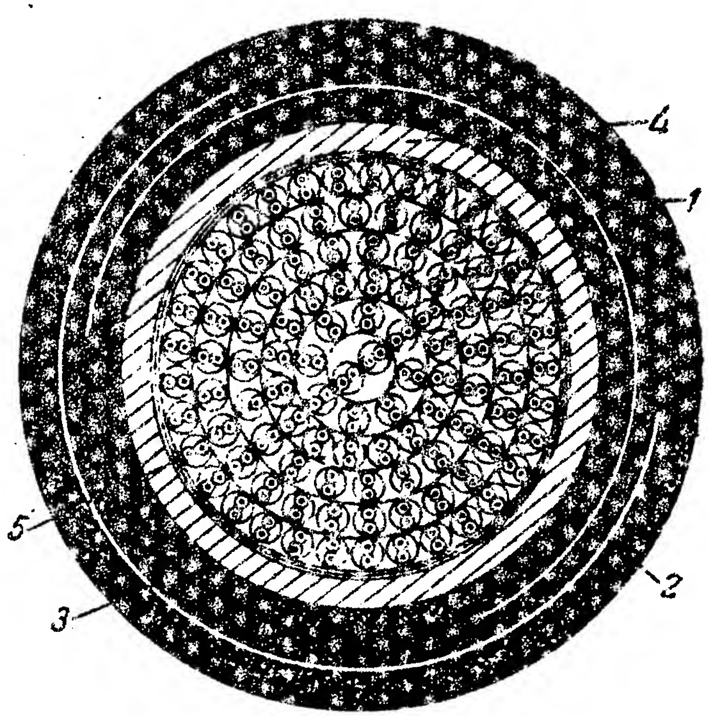


Fig. 3-4. Type TB urban telephone cable.
1--outer protective coating; 2--armor consisting of two iron strips; 3--bedding; 4--lead sheathing; 5--insulated conductors

b) Telephone distribution cables (Type TRK)

These are designed for leading-in terminal blocks in cabinets and distribution boxes, and for laying along the outside and inside walls of buildings.

The current-carrying conductors are made of enameled annealed copper wire 0.5 mm in diameter and are wound with cotton thread; a double winding of cotton thread may be used without the enamel. Contrasting colors are provided for the conductors. The two conductors are twisted (with the

Table 3-5

(1) Разбивка пар по повивам в городских телефонных кабелях

(2) Число пар			(3) Номера повивов																		
Номинальное (4)	Фактическое (5)	Центральные (6)	1	2	3	4	5	6	7	8	9	10	11	12	13	14	16	16	17	13	19
5	5	0	5	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
10	10	2	8	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
20	20	1	6	13	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
30	30	4	10	16	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
40	40	1	7	13	19	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
50	50	4	10	15	21	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
70	70	2	8	14	20	26	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
80	80	4	10	16	22	28	—	—	—	—	—	—	—	—	—	—	—	—	—	—	—
100	101	2	8	14	20	26	31	—	—	—	—	—	—	—	—	—	—	—	—	—	—
150	151	4	10	16	22	28	33	38	—	—	—	—	—	—	—	—	—	—	—	—	—
200	201	4	10	16	22	28	34	40	47	—	—	—	—	—	—	—	—	—	—	—	—
300	302	8	15	21	28	34	40	46	52	58	—	—	—	—	—	—	—	—	—	—	—
400	402	1	7	13	19	25	31	37	43	48	54	60	64	—	—	—	—	—	—	—	—
500	503	3	9	15	21	27	33	39	45	51	57	62	67	74	—	—	—	—	—	—	—
600	603	5	11	17	23	29	35	40	46	52	58	64	69	74	80	—	—	—	—	—	—
700	704	1	6	12	17	23	29	35	41	47	53	59	65	70	76	83	85	—	—	—	—
800	804	6	12	18	24	30	36	42	48	54	59	65	70	76	82	88	94	—	—	—	—
900	905	6	12	18	24	30	36	42	48	64	59	65	70	76	82	88	94	101	—	—	—
1 000	1 005	6	12	17	23	29	36	41	47	53	59	65	71	77	82	88	94	100	106	—	—
1 200	1 205	4	10	16	22	28	34	40	46	52	58	64	70	76	82	87	93	99	109	109	111

1. BREAKDOWN OF LAYS BY PAIRS IN URBAN TELEPHONE CABLES
2. Number of pairs
3. Number of layers
4. Rated (nominal)
5. Actual
6. Central

Fig. 3-5. Telephone distribution cable of Type TRK (single-pair).
1--lead sheathing; 2--cotton-thread winding; 3--insulated conductors.

exception of the single-pair type) into a pair and wound with a cotton-thread spiral. The pairs are regularly twisted and calico stripping or a double paper stripping is wound, layer by layer, into the cable. All insulation is impregnated with an insulating compound. The lead sheathing is applied over the insulation.

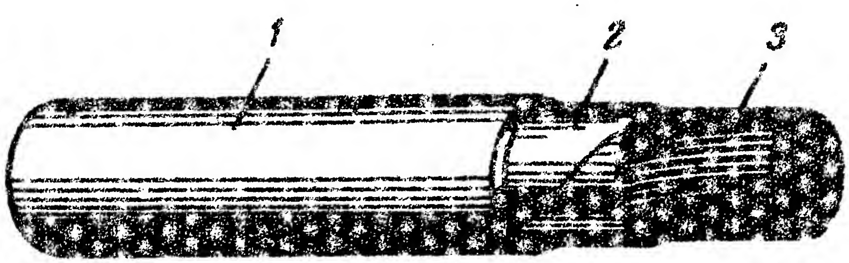


Fig. 3-6. Type TRK (multi-pair) telephone distribution cable.
1--lead sheath; 2--winding of paper or cloth stripping; 3--insulated pair of conductors twisted in concentric layers

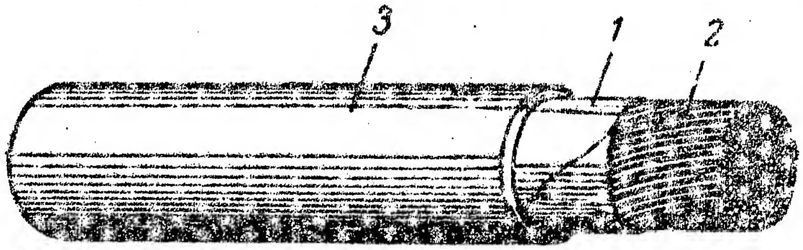


Fig. 3-7. Type TSS Telephone office cable.
1---Winding of paper or cloth stripping; 2--insulated conductors; 3--lead shield.

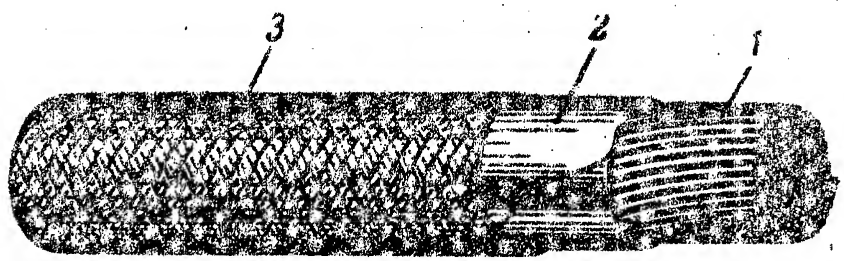


Fig. 3-8. Type TSO braided telephone office cable.
1--Insulated conductors; 2--Double winding of varnished cambric or oiled paper; 3--braiding of twisted cotton thread impregnated with moisture-resistant compound.

Fig. 3-5 and 3-6 show single-pair and 30-pair distribution cables.

The basic dimensions of Type TRK cables are given in Table 3-7. The thickness of the lead sheathing is given at the end of the Chapter.

c) Station telephone cables.

These cables are used in the installation of telephone stations.

Station cables are classified, in accordance with

Table 3-6

Основные размеры городских телефонных каб

Парность кабеля	Диаметр жилы 0,5 мм. Наружный диаметр, мм: ТГ	ТА	ТБГ	ТБ	ТП	ТК	Максимальная строительная длина, м: ТГ	ТБГ	ТБ	Диаметр жилы 0,6 мм. Наружный диаметр, мм: ТГ	ТА	ТБГ	ТБ	ТП	ТК	Максимальная строительная длина, м: ТГ и ТА	ТБГ, ТБ и ТП	ТК	ТГ
5	8	12	13	16	18	—	1 000	1 000	—	9	13	14	17	19	—	1 000	1 000	—	425
10	9	13	14	17	19	—	1 000	1 000	—	11	15	16	19	21	—	1 000	1 000	—	590
20	11	15	16	19	21	25	1 000	1 000	1 000	14	18	19	22	24	28	1 000	1 000	1 000	790
30	13	17	18	21	23	27	1 000	1 000	1 000	17	21	22	25	27	31	1 000	1 000	1 000	1 050
40	14	18	19	22	24	28	1 000	1 000	1 000	19	23	24	27	29	33	1 000	1 000	1 000	1 220
50	16	20	21	24	26	30	1 000	1 000	1 000	21	25	26	29	31	35	1 000	1 000	800	1 460
70	18	22	23	25	28	32	1 000	1 000	800	23	27	28	31	33	37	1 000	800	600	1 690
80	19	23	24	27	29	33	1 000	1 000	700	24	29	30	34	35	44	1 000	700	500	1 910
100	20	24	25	28	30	34	1 000	1 000	500	27	32	33	37	38	47	1 000	600	400	2 170
150	24	29	30	34	35	44	1 000	600	450	32	37	38	42	43	52	600	500	350	3 020
200	27	32	33	37	38	47	1 000	600	350	37	42	43	47	48	57	400	400	300	3 890
300	30	35	36	40	41	50	800	400	300	44	49	50	54	56	64	350	300	225	5 330
400	37	42	43	47	48	57	500	350	200	50	55	58	60	62	70	300	250	150	6 760
500	42	47	48	52	54	62	400	300	200	56	61	62	66	68	76	250	200	150	8 270
600	45	50	51	55	57	65	360	250	150	61	66	67	71	73	81	200	150	150	9 560
700	49	54	—	—	—	—	300	—	—	66	71	—	—	—	—	150	—	—	10 810
800	52	57	—	—	—	—	250	—	—	70	75	—	—	—	—	150	—	—	11 840
900	55	60	—	—	—	—	250	—	—	73	78	—	—	—	—	150	—	—	12 350
1 000	58	69	—	—	—	—	250	—	—	75	80	—	—	—	—	150	—	—	13 860
1 200	63	68	—	—	—	—	200	—	—										

Table 3-6
(Continued)

елей

Диаметр жилы 0,7 мм 7								
Наружный диаметр, мм 4						Максимальная строительная длина, м 5		
ТГ 8	ТА 14	ТГБ 11	ТБ 12	ТП 13	ТК 15	ТГ, ТА 16	ТГТ, ТБ и ТП 17	ТК 15
10	14	15	18	20	—	1 000	1 000	—
12	16	17	20	22	—	1 000	1 000	—
16	20	21	24	26	30	1 000	1 000	1 000
20	24	25	28	30	34	1 000	1 000	1 000
22	26	27	30	32	36	1 000	1 000	1 000
24	29	30	34	35	44	1 000	1 000	1 000
28	33	34	38	39	48	1 000	800	800
30	35	36	40	41	50	1 000	700	600
33	38	39	43	44	53	1 000	600	500
39	44	45	49	51	59	600	500	400
45	50	51	56	57	65	400	400	350
54	59	60	64	66	74	300	300	300
62	67	68	72	74	32	250	250	200
69	74	75	79	81	89	200	200	150
75	30	81	85	87	95	150	150	150

[Key to Table 3-6:]

1. BASIC DIMENSIONS OF URBAN TELEPHONE CABLES
2. Pair count of cable
3. Conductor diameter 0.5mm
4. Outside diameter, mm
5. Greatest shipping length, m
6. Conductor diameter 0.6mm
7. Conductor diameter 0.7mm
8. TG
9. TG and TA
10. TBG, TB, and TP
11. TBG
12. TB
13. TP
14. TA
15. TK
16. TG, TA
17. TGT, TB, TP

the type of protective covering used, into lead-clad (Type TSS) cables, which are laid on the outside of buildings and in premises with elevated humidity (see Fig. 3-7), and braided cables (Type TSO), which are used in dry environments (see Fig. 3-8).

The current-carrying conductors are made of enameled annealed copper wire 0.5mm in diameter with a cotton-thread winding.

The differently colored conductors are twisted into groups of 2, 3, or 4, and these are wound with a spiral of

cotton thread and then twisted properly (by layers) into a cable.

The overall twist is wound with a cambric or double paper strip and, in TSO cables, with two layers of oiled paper and varnished-cambric stripping. In this form, the

Table 3-7

BASIC DIMENSIONS OF TELEPHONE DISTRIBUTION CABLES
(Type TRK)

(2) Число пар	Максимальный наружный (3) диаметр кабеля, мм	Максимальная строительная (4) длина, м
1	3,5×4,5	1 000
5	8,0	1 000
10	10,5	1 000
20	13,5	1 000
30	15,5	1 000
40	17,0	1 000
50	19,5	1 000
70	22,5	1 000
100	26,0	1 000

2. Number of pairs
3. Maximum outside diameter of cable, mm
4. Maximum shipping length, m

cable is impregnated with an insulating composition; then TSS cables are covered with the lead sheath and TSO cables with cotton-thread braiding which has been colored with an oil-base dye.

The production of station cables with an outer sheath of polyvinyl-chloride plastic instead of lead has recently been initiated. These are cables of Type TSSh, where the

last letter indicates that the cable has a vinyl jacket (shlang) instead of a lead one.

The basic dimensions of the station cables are given in Table 3-8.

The station cables have a lead sheath 0.9mm thick.

Table 3-8

(1) Основные размеры станционных кабелей

(2) Система скрутки (число и состав групп)	Максимальный наружный диаметр, мм (3)		Максимальная строительная длина, м (6)
	ТСС (4)	ТСО (5)	
5×3	10,0	9,5	1 000
11×3	13,0	12,5	1 000
21×3	15,0	14,5	1 000
11×4	15,0	14,5	1 000
21×3	16,0	15,5	1 000
26×3	18,0	18,0	1 000
63×3	25,0	25,0	1 000
103×3	30,0	30,0	1 000

1. BASIC DIMENSIONS OF STATION CABLES
2. Twisting system (number and composition of groups)
3. Maximum outer diameter (mm)
4. TSS
5. TSO
6. Maximum shipping length, m

Station telephone cables with somewhat different construction are encountered. For example, we may have cable with tinned copper wires, which make brazing easier, or with silk insulation or layers of silk and cotton thread.

The practice of varnishing the surface of the conductor insulation has recently come into wide use; this makes them easier to work with in assembling station equipment. Distribution and station cables are made with polyvinyl chloride conductor insulation and protective sheathing.

d) Cables for signaling and holding

These cables are employed in railroad remote control (STsB), telegraph, and other systems.

Cables with protective coverings appropriate to the conditions of use are empolyed.

The current conductors are of annealed copper wire 1 mm in diameter. The conductors, insulated with a solid layer of 5-6 strips of cable paper, are twisted with a packing of cable thread or with a paper plait. The twisted cable is wound on the outside with cambric or paper stripping and then impregnated with insulating compound.

A lead sheath and, in armored cables, a protective covering, are applied over the winding. Their thicknesses are listed at the end of the present chapter.

The basic dimensions of signaling and holding cables are given in Table 3-9.

e) Long-distance communications cables (pack-thread cables)

A general picture of the design of the basic elements going into packthread cables was given at the beginning of the present chapter (Tables 3-1 and 3-2). The number of

Table 3-9

(1) **Основные размеры некоторых кабелей для сигнализации и блокировки**

(2) Число жил	(3) Приблизительный наружный диаметр, мм			Максимальная строительная длина, м		
	(4) для марки СОГ	(5) для марок СОА, СОБ, СОБГ, СОП и СОПГ	(6) для марки СОК	(4) для марки СОГ	(5) для марок СОА, СОБ, СОБГ, СОП, СОПГ	(6) для марки СОК
3	9,0	15,0	—	1 000	1 000	—
5	10,0	16,0	—	1 000	1 000	—
7	11,0	17,0	—	750	750	—
12	13,5	20,0	30,0	750	750	750
19	15,5	22,0	32,0	750	750	500

1. BASIC DIMENSIONS OF CERTAIN SIGNALING AND HOLDING CABLES
2. Number of pairs
3. Approximate outside diameter, mm
4. for Type SOG
5. for Types SOA, SOB, SOBG, SOP, SOPG
6. for Type SOK

elements and their combinations are determined by the natrue and number of communications circuits to be transmitted via the cable in question. The long-distance cables used in this country may be classed into three consolidated

groups: 1) simple and screened low-frequency cables with the quadded structure for audio-frequency transmission, 2) composite low-frequency cables, and 3) high-frequency cables for the 12-channel system (to 60 kcps) and the 24-channel system (to 108 kcps).

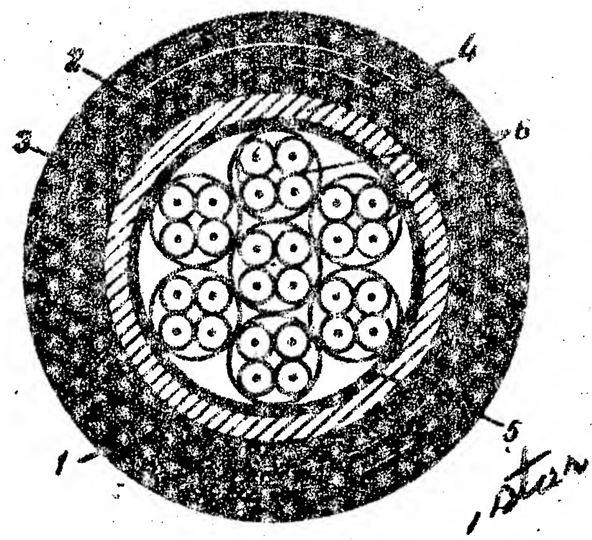


Fig. 3-9. Type TZB spiral-four Com-cunications cable.

1--protective outer covering; 2--armor formed from two iron strips; 3--bedding; 4--lead shield; 5--paper-tape winding; 6--insulated quadded conductors.

Low-frequency quad cables are used as connecting lines between district (rayon) automatic telephone offices (ATS), for river crossings, and for cabling telephone-telegraph centers. They consist of a certain number of quads twisted together, with the quads screened from one another in certain cases (usually every other one). Fig.

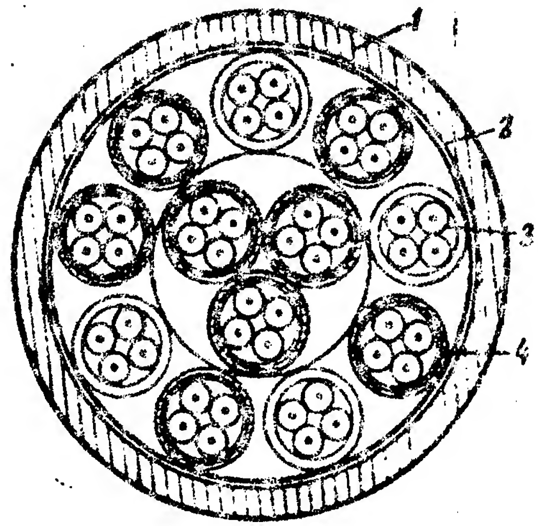


Fig. 3-10. Type TZEG screened quadded communications cable.
1.--lead sheathing; 2--paper-tape winding; 3--unscreened quadded conductors; 4--screened quadded conductors.

Table 3-10

(1) Основные данные кордельных кабелей

Марка кабеля (3)	(6) Диаметр проволоки, мм		
	0,8 и 0,9	1,0 и 1,2	1,4
	(7) Число четверок		
ТЗГ, ТЗБ (2)	3, 4, 7, 12, 19, 24, 27, 30, 37, 44, 48, 52, 61, 75, 80, 91, 102, 108, 114	3, 4, 7, 12, 14, 19, 24, 27, 30, 34, 37, 48, 52, 61	3, 4, 7, 12, 14, 19, 24, 27, 30, 37
ТЗЭГ, ТЗЭБ, ТЗБГ, ТЗЭБГ, ТЗП, ТЗЭП, ТЗПГ, ТЗЭПГ (4)	3, 4, 7, 12, 14, 19, 24, 27, 30, 37,	3, 4, 7, 12, 14, 19, 24, 27, 30, 37	3, 4, 7, 12, 14
ТЗК, ТЗЭК (5)	7, 12, 14, 19, 24, 27, 30, 37	3, 4, 7, 12, 14, 19, 24, 27, 30, 37	3, 4, 7, 12, 14

[Key to Table 3-10]

1. BASIC DATA FOR PACKTHREAD CABLES
2. TZG, TZB
3. Type of cable
4. TZEG, TZEB,
   TZBG, TZEBG,
   TZP, TZEP,
   TZPG, TZEPG,
5. TZK, TZEK
6. Diameter of wire, mm
7. Number of quads

[Key to Table 3-11]

1. OUTSIDE DIAMETERS OF TYPE TZG, TZB, TZP, AND TZK CABLES
2. Number of quads
3. TZG
4. Diameter of wire, mm
5. TZB
6. TZP
7. TZK

[Key to Table 3-12] (pages following Table 3-11)

1. TYPES OF SINGLE-LAY COMPOSITE CABLES
2. Sectional vies in (conventional) schematic representation
3. Designation (conventional)
4. Approximate diameter of cable under lead sheath, mm
5. ekr [= screened]
6. Type [I, III, etc.]
7. Table 3-12 (continued)
8. u [=reinforced]

3-9 and 3-10 show sections of simple and screened quadded cables. The numbers of quads from which they are twisted are given in Table 3-10. The same table also indicates the diameters of the current-conducting

Table 3-11

1 Наружный диаметр кабелей марок ТЗГ, ТЗБ, ТЗП, ТЗК

2 Число четверок	3 ТЗГ					5 ТЗБ					6 ТЗП				
	4 Диаметр проволоки, мм					4 Диаметр проволоки, мм					4 Диаметр проволоки, мм				
	0,8	0,9	1,0	1,2	1,4	0,8	0,9	1,0	1,2	1,4	0,8	0,9	1,0	1,2	1,4
3	12,5	13	13,5	14,5	18	21,5	22	22,5	22,5	26	—	24,5	25	23,0	20,5
4	13,5	14	14,5	16	20	22,5	23	23,5	24	28	24,5	25	25,5	25	31
7	16	16,5	17,6	19	24	25	26,5	26	27	33	27	27,5	28	28	35
12	21,5	22	23,5	25,5	32,5	31,5	32	32,5	34,5	41,5	32,5	33	33,5	36,5	44
14	22,5	23,5	25	27	34,5	33	33,5	34	35,5	43	34	34,5	36	36,5	45
19	25	26	27,5	30	38,5	36,5	37	37,5	39	47	—	—	—	—	—
27	30,5	32	33,5	36,5	46,5	41	41,5	42	45	55,5	—	—	—	—	—
30	31,5	33	35	38	48,5	42	42,5	45	46,5	57,5	—	—	—	—	—
37	34	36,5	37,5	41	53	46	46,5	47	49,5	61,5	—	—	—	—	—
44	—	40	—	—	—	—	48,5	—	—	—	—	—	—	—	—
48	—	41	—	—	—	—	49,5	—	—	—	—	—	—	—	—
51	—	42	—	—	—	—	50,5	—	—	—	—	—	—	—	—
61	—	46	—	—	—	—	54,5	—	—	—	—	—	—	—	—
75	—	50,5	—	—	—	—	58,5	—	—	—	—	—	—	—	—
89	—	51,5	—	—	—	—	60,5	—	—	—	—	—	—	—	—
91	—	54	—	—	—	—	62,5	—	—	—	—	—	—	—	—
103	—	60	—	—	—	—	67,5	—	—	—	—	—	—	—	—
114	—	61	—	—	—	—	69	—	—	—	—	—	—	—	—

Table 3-11
Continued

(7) ТЗК				
(4) Диаметр проволоки, мм				
0,8	0,9	1,0	1,2	1,4
—	—	—	31	35
—	—	31	32	37
32,5	33,5	34	35,5	41
—	38,5	—	—	—
—	39,5	—	—	—
—	42	—	—	—
—	49,5	—	—	—
—	50,5	—	—	—
—	56	—	—	—
—	62	—	—	—
—	63	—	—	—
—	64	—	—	—
—	67,1	—	—	—
—	72	—	—	—
—	73	—	—	—
—	75	—	—	—
—	82	—	—	—
—	83	—	—	—

Table 3-12

1 Типы одноосновных комбинированных кабелей

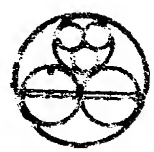
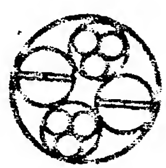
Разрез (условно) схематического изображения 2	3 Обозначение (условно)	Приблизительный диаметр кабеля по свинцовой оболочке, мм 4
 6 Тип I	5 1×2 экр.×1,2 + 2× (3×2×0,8) 1×2 экр.×1,4 + 2× (3×2×0,8)	17 17,5
 Тип II	2×2 экр.×1,2 + 1× (3×2×0,8) 2×2 экр.×1,4 + 1× (3×2×0,8)	17 17,5
 Тип III	2×2 экр.×1,2 + 2 ×(3×2×0,8) 2×2 экр.×1,4 + 2× (3×2×0,8)	19 19,5

Table 3-12
(Cont'd)

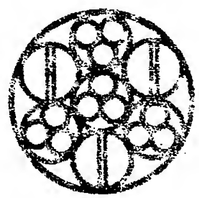
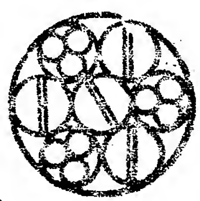
 Тип IV	3×2 экр.×1,2 + 1× (3×2×0,8) 3×2 экр.×1,4 + 1× (3×2×0,8)	19 19,5
 Тип V	3×2 экр.×1,2 + 4× (3×2×0,8) 3×2 экр.×1,4 + 4× (3×2×0,3)	23,5 24
 Тип VI	4×2 экр.×1,2 + 3× (3×2×0,8) 4×2 экр.×1,4 + 3× (3×2×0,8)	23,5 24

Table 3-12 /Continued

Продолжение табл. 3-12

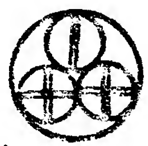
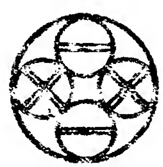
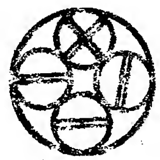
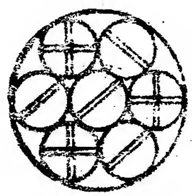
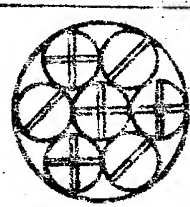
Разрез (условно) схематического изображения	Обозначение (условно)	Приблизительный диаметр кабеля под свинцовой оболочкой *мм*
 Тип VII	1×2 экр.×0,9 + 2× (1×4у×0,8)	12,5
 Тип VIII	2×2 экр.×0,9 + 1× (1×4у×0,8)	12,5
 Тип IX	2×2 экр.×0,9 + 2× (1×4у×0,8)	13,5

Table 3-12
Continued

 Тип X	3×2 экр.×0,9 + 1× (1×4у×0,8)	13,5
 Тип XI	4×2 экр.×0,9 + 3× (1×4у×0,8)	17
 Тип XII	3×2 экр.×0,9 + 4× (1×4у×0,8)	17

wires used in them. The cables are twisted by regular concentric layers according to the system indicated in Table 3-3.

The thicknesses of the lead sheathings and the protective coverings are listed at the end of the chapter. The basic dimensions of the unscreened cables are presented in Table 3-11.

The 1949 Standard sets forth strictly determined shipping lengths for these cables; they must be 425 m or multiples thereof (850m, 1275m) in length.

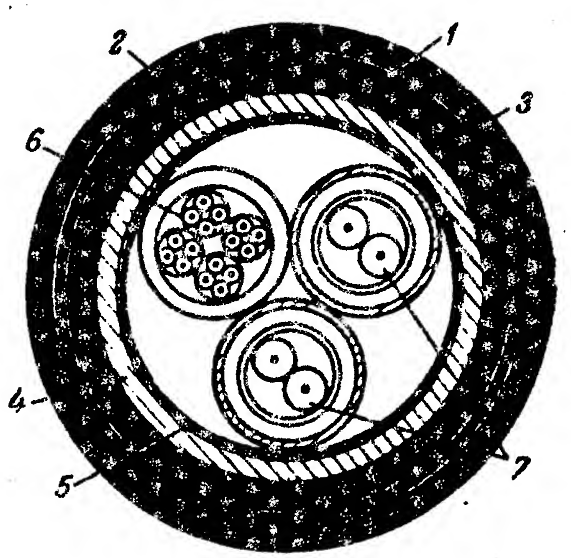


Fig. 3-11. Type TDSB composite communications cable.
1--outer covering; 2--armor formed by two iron strips; 3--bedding; 4--lead sheath; 5--paper-tape winding; 6--unscreened quads; 7--screened pairs.

Composite low-frequency cables are used for inter-urban telephone, telegraph, facsimile, and in certain cases

even broadcast (music) transmissions.

Cables of this type are unified under a common Standard (ST-5-4), which classifies them into 1- and 2-layer types and provides for all possible combinations of the normal elements.

Single-layer cables are composed either of screened pairs with conductors of 1.2-and 1.4-mm wire in combination with reinforced sextuplets of equal diameter with conductors made from wire 0.8mm in diameter, or of screened pairs with conductors 0.9mm in diameter in combination with reinforced quads of equal diameter with conductors of 0.8-mm wire.

Here combinations of 3, 4, and 7 such elements can be made up. Al combinations that have found practical use reduce to the twelve basic types of cables listed in Table 3-12. Representations of these appear in Figs. 3-11 and 3-12. The design of these cables is distinguished by the fact that in them, individual conductors made of wires of different diameters are combined into elements of approximately equal strength. This excludes the possibility of nonsymmetrical distribution of mechanical stresses, which can result in breakage of the conductors at bends and especially during laying of the cable.

Two-layer composite cables contain screened pairs

with wire diameters of 0.9, 1.0, 1.2 or 1.4 mm in their central (first) layer, and spiral fours or pairs with wire diameters of 0.7, 0.8, and 0.9 mm in the outer layer. All possible useful group-formations are shown in Table 3-13.

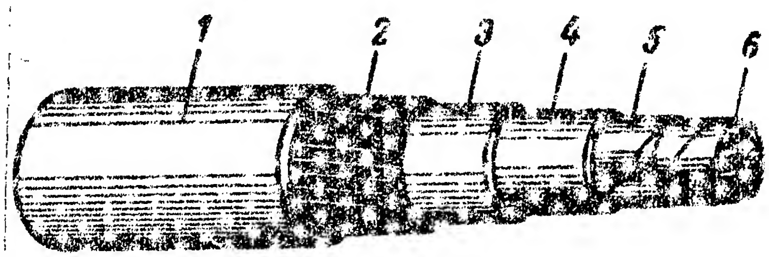


Fig. 3-12. Type TDSP composite communications cable.

1--Outer jute covering; 2--armor of flat steel wire; 3--Jute bedding; 4--lead sheath; 5--paper winding; 6--various elements laid up into cable.

The thicknesses of the lead sheaths and protective coverings are listed at the end of the chapter.

High-frequency cables (for the 12-channel system) differ little in structure from the low-frequency types described above. However, they are prepared from more stable elements in order to obtain improved electrical parameters. For this purpose, the conductors of these cables are insulated with packthread with a shorter lay and by two layers of paper tape; in addition, such technological measures as special selection of the packthread, the use of paper with the highest possible air permeability

Table 3-13

1 Типы двухповивных кабелей

2 Число элементов (экр. пар) и диаметр проволоки жил в центре, мм	3 Число элементов (n) во внешнем (втором) повиве и приблизительный диаметр (D) кабелей под свинцовой оболочкой											
	1×4×0,7		1×4×0,8		1×4×0,9		1×2×0,7		1×2×0,8		1×2×0,9	
	4 n	5 D, мм	n	D, мм	n	D, мм	n	D, мм	n	D, мм	n	D, мм
1×2 экр.×0,9	8	15	7	15,5	7	16	9	14,5	9	15	9	15,5
1×2 экр.×1,0	9	16,5	8	17	8	17,5	10	15,5	10	16	10	16,5
1×2 экр.×1,2	9	17	9	17,5	9	18	11	16	11	17	10	17,5
1×2 экр.×1,4	9	17,5	9	18	9	18,5	11	17	11	17,5	10	18
2×2 экр.×0,9	12	21	11	21,5	11	22	11	20	13	20,5	13	21
2×2 экр.×1,0	14	23	13	23,5	13	24	16	22,5	15	23	15	23,5
2×2 экр.×1,2	15	24,5	14	25	14	25,5	17	24	16	24,5	16	25
2×2 экр.×1,4	15	25,5	15	26	14	26,5	18	25	17	25,5	16	26
3×2 экр.×0,9	12	21,5	12	22	12	23,5	14	21	14	21,5	13	22
3×2 экр.×1,0	14	24	14	24,5	13	25	17	23,5	16	24	15	24,5
3×2 экр.×1,2	15	25	15	25,5	14	26	17	24,5	17	25	16	25,5
3×2 экр.×1,4	16	26	15	27	15	27	18	25,5	18	26	17	26,5
4×2 экр.×0,9	12	21,5	12	22,5	12	23	17	21,5	16	22	15	23
4×2 экр.×1,0	15	24,5	14	25	14	25,5	20	24	19	24,5	18	25
4×2 экр.×1,2	16	25,5	15	26	15	26,5	21	25	20	25,0	19	26
4×2 экр.×1,4	17	26,5	16	27,5	15	27,5	22	26	21	26,5	20	27

Key to Table 3-13

1. TYPES OF TWO-LAYER CABLES
2. Number of elements (screened pairs) and wire diameter of conductors in center, mm
3. Number of elements (n) in external (second) layer and approximate diamater (D) of cables under lead sheathing
4. n
5. D, mm

(to lower the dielectric strength), selective calibration of the wire, careful control and monitoring of the tension on the conductors during twisting, etc., are taken. In some cases, paper packthread and tape are replaced by styroflex materials. Further, a special packthread filler is placed between the insulated conductors within the quads in order to stabilize the compling coefficients of the high-frequency cables.

Several types of high-frequency cables with packthread-paper and styroflex insulation have recently gone into production.

As an example, Fig. 3-13 shows a schematic representation of the structure of a 32-pair composite cable for the 12-channel system (up to 60 kcps) of Type MKB-32x2. A single screened pair with enameled 0.9-mm copper wire is placed at the center of this cable. In the first layer (from the center) we have three screened pairs with conductors of wire 1.4mm in diameter and two reinforced quads with conductors of wire 1.2mm in diameter. The first layer is separated from the second by a winding of four layers of Type K-12 cable paper. The second (outer) layer is laid up from twelve spiral fours with conductors of 1.2-mm wire and wound with four layers of K-12 cable paper. The cable is leaded and armored by two 45x0.5-mm steel strips with

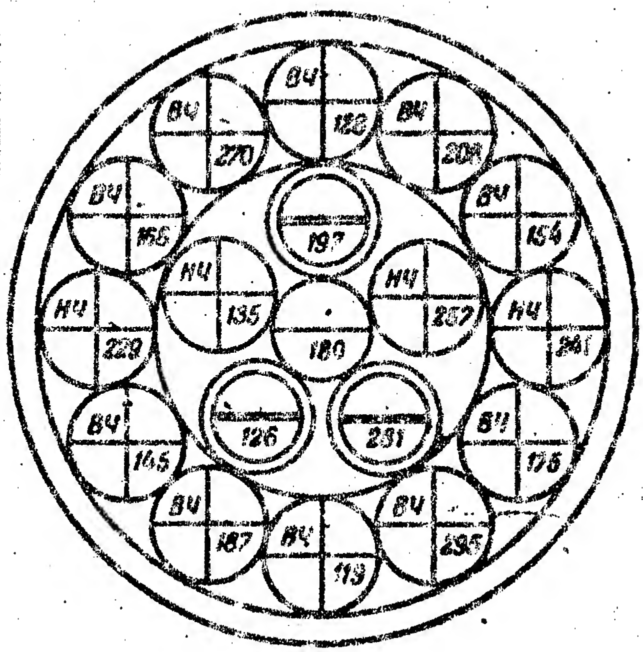


Fig. 3-13. Structure of symmetrical 32x2 cable.

1--The lay of the first layer is 400mm (right lay).

2--The lay of the second layer is 680mm (left lay).

3--The lays of the group are indicated on the drawing.

4-- ВЧ = high-frequency quad.

5-- НЧ = low-frequency quad.

an outer covering of impregnated jute which has been soaked with a bitumen preparation. The lead sheath is 2.0mm thick. The overall weight of the cable is 6270 kg/km.

3-3. FUNDAMENTALS OF DESIGN CALCULATION FOR SYMMETRICAL CABLES

The design description that has been adopted in our cable industry consists of two sections: the first lists the sectional dimensions of all design elements of the cable in millimeters, and the second the outlay of production materials used in their fabrication, expressed in kilograms per kilometer of cable length. The design description, which is called "Design" for short, is the basic technical document used as a guide in the fabrication, planning and accounting aspects of the cable industry. It is similar in purpose to the blueprint used in machine construction. The methodology of the Scientific Research Institute of the Cable Industry (NIIKP)--the basic points of which regarding the design of communications cables are presented below--is used in the development of a cable product.

General Specifications

a) In the calculations, the geometrical dimensions of the component elements of the cable are those without consideration of tolerances (i.e., the nominal dimensions).

The outlay of all materials is computed from theory, without consideration of tolerances and waste, etc., in kg

per km of cable (i.e., the net weight P).

b) The final weight data for the outlay of materials are indicated on the designs with an accuracy to 1.01 kg for weights to 10 kg, to 0.1 kg for a weight of 1 kg, and to 1 kg for weights above 100 kg.

c) The specific weights adopted for the materials are:

Copper	8.89
Steel (armoring band and wire)	7.8
Lead	11.4
Polyvinyl chloride plastic	1.32--1.52
Polyethylene	0.95
Telephone paper	0.8
Paper pulp	0.75
Paper packthread	0.85--0.95
Cable paper	0.83

d) The thicknesses of the lead sheathings are given in Tables 3-14, 3-15, 3-16, and 3-17.

e) The bedding of armored cable consists of the following, applied in succession:

1) A layer of binding compound (a bitumen preparation), 2) two layers of previously impregnated cable paper, 3) a layer of binding compound, 4) a layer of previously impregnated cable thread (jute), 5) a layer of binding compound.

A bedding of 6 layers of cable paper without cable thread is admissible for urban telephone cables and for quadded communications cables with packthread-paper

insulation (with the exception of Types TZK, TZEK, and cables for CTsB).

Table 3-14

Толщина свинцовой оболочки городских телефонных кабелей в зависимости от диаметра и марки кабеля (постановление Гостехники СССР № 326 от 2/IV 1948 г.) 1

2 Диаметр кабеля под свинцовой оболочкой, *мм*	3 Толщина свинцовой оболочки кабеля марки					
	4 ТГ		5 ТА, ТБ, ТБГ и ТП		6 ТК	
	минимальная, *мм* 7	номинальная, *мм* 8	минимальная, *мм* 7	номинальная, *мм* 8	минимальная, *мм* 7	номинальная, *мм* 8
9 До 6	1,0	1,15	1,0	1,15	—	—
До 8	1,1	1,25	1,0	1,15	—	—
До 13	1,2	1,4	1,1	1,25	1,8	2,05
До 16	1,3	1,5	1,2	1,4	1,8	2,05
До 20	1,4	1,6	1,3	1,5	1,9	2,15
До 23	1,5	1,7	1,3	1,5	2,0	2,3
До 26	1,6	1,8	1,4	1,6	2,0	2,3
До 30	1,7	1,95	1,5	1,7	2,1	2,4
До 33	1,8	2,05	1,6	1,8	2,1	2,4
До 36	1,9	2,15	1,6	1,8	2,2	2,5
До 40	2,0	2,3	1,8	2,05	2,2	2,5
До 43	2,1	2,4	1,8	2,05	2,3	2,6
До 46	2,2	2,5	1,9	2,15	2,4	2,7
До 50	2,3	2,6	2,0	2,3	2,5	2,8
До 53	2,4	2,7	2,0	2,3	2,5	2,8
До 56	2,5	2,8	2,1	2,4	2,6	2,9
До 60	2,6	2,9	2,2	2,5	2,7	3,0
10 Свыше 60	2,7	3,0	2,3	2,6	2,8	3,1

1. THICKNESS OF LEAD SHEATHS OF URBAN TELEPHONE CABLES AS A FUNCTION OF THE DIAMETER AND TYPE OF THE CABLE (SET FORTH BY GOSTEKHNIKA USSR No. 326 of 2 April 1948).
2. Diameter of cable under lead sheathing, mm
3. Thickness of lead sheathing of cable of Type [see 4,5,6]
4. TG
5. TA, TB, TBG, and TP
6. TK
7. Minimum, mm
8. Nominal, mm
9. To
10. Above

Table 3-16

Толщина свинцовой оболочки кордельных кабелей (ГОСТ 5008-49)

Диаметр кабеля под свинцовой оболочкой, мм	Радиальная толщина свинцовой оболочки в мм для кабелей: бронированных железными лентами или плоскими проволоками		голых освинцованных		бронированных круглыми стальными проволоками	
	минимальная	номинальная	минимальная	номинальная	минимальная	номинальная
До 13	1,1	1,25	1,2	1,4	1,8	2,05
Свыше 13 до 16	1,2	1,4	1,3	1,5	1,8	2,05
" 16 " 20	1,3	1,5	1,4	1,6	1,9	2,15
" 20 " 23	1,3	1,5	1,5	1,7	2	2,3
" 23 " 26	1,4	1,6	1,6	1,8	2	2,3
" 26 " 30	1,5	1,7	1,7	1,95	2,1	2,4
" 30 " 33	1,6	1,8	1,8	2,05	2,1	2,4
" 33 " 36	1,6	1,8	1,9	2,15	2,2	2,5
" 36 " 40	1,8	2,05	2,0	2,3	2,2	2,5
" 40 " 43	1,8	2,05	2,1	2,4	2,3	2,6
" 43 " 46	1,9	2,15	2,2	2,5	2,4	2,7
" 46 " 50	2	2,3	2,3	2,6	2,5	2,8
" 50 " 53	2	2,3	2,4	2,7	2,5	2,8
" 53 " 56	2,1	2,4	2,5	2,8	2,6	2,9
" 56	2,2	2,5	2,6	2,9	2,7	3,0

1. THICKNESS OF LEAD SHIELDING ON PACKTHREAD CABLES (GOST 5008-49)
2. Diameter of cable under lead sheathing, mm
3. Radial thickness of lead sheath in mm for cables [as follows]:
4. Armored with iron band or flat wire
5. Bare lead-covered
6. Armored with round steel wire
7. Minimum
8. Nominal
9. To
10. Above (from)

Table 3-16

1 Толщины свинцовой оболочки кабеля для сигнализации и блокировки (постановление Гостехники СССР № 323 от 2/VI 1948 г.)

Диаметр кабеля под свинцовой оболочкой, мм 2	3 Толщина оболочки для кабелей марок, мм			
	4 СОГ, СОА	5 СОБ, СОБГ, СОП, СОПГ	6 СОК	
	7 минимальная	8 номинальная	7 минимальная	8 номинальная
9 До 13	0,9	1,05	1,2	1,4
До 16	1,0	1,15	1,3	1,5
До 20	1,1	1,25	1,4	1,6
До 23	1,2	1,4	1,5	1,7

1. THICKNESSES OF LEAD SHEATHING ON CABLE FOR HOLDING AND SIGNALING (AS SET FORTH BY GOSTEKHNIKA USSR No. 323 of 2 April 1948)
2. Diameter of cable under lead sheathing, mm
3. Thickness of sheathing for cable of Types as follows, mm
4. SOG, SOA
5. SOB, SOBG, SOP, SOPG
6. SOK
7. Minimum
8. Nominal
9. To

The thickness of the covering and the nature of the armor on various types of cables are indicated in Tables 3-18, 3-19 and 3-20.

The thickness of the cable-paper bedding is taken as 2mm for all cable sizes. The outer covering of armored cables of all types is formed by successively applied layers as follows: binding compound, preimpregnated thread, binding compound, and chalk solution.

Table 3-17

(1) Минимальная толщина свинцовых оболочек телефонных распределительных кабелей (ВТУЭ-316-43)

(2) Число пар	Минимальная толщина свинцовой оболочки, *мм* (3)
1	0,8
5	0,8
10	0,8
20	0,9
30	0,9
40	0,9
50	0,9
70	1,0
100	1,0

1. MINIMUM THICKNESS OF LEAD SHEATHING ON TELEPHONE DISTRIBUTION CABLES (VTUE-316-43).
2. Number of pairs
3. Minimum thickness of lead sheathing, mm

Table 3-18

(1) Толщина защитных покровов городских телефонных кабелей (постановление Гостехники СССР № 327)

аметр кабеля поверх свинцовой оболочки, *мм* (2)	(3) Номинальная толщина защитных покровов, *мм*				
	(4) Подушка	(5) Броня из			(6) Наружны покров
		стальных лент (7)	(8) оцинкованных стальных проволок		
			(9) круглых	(10) плоских	
До 13	1,5	2×0,3	—	—	1,5
До 23	1,5	2×0,5	1,5	4	2,0
До 37	2,0	2×0,5	1,5	4—6	2,0
До 50	2,0	2×0,5	1,7	6	2,0
(11) 50 и выше	2,5	2×0,8	1,7	6	2,0

[Key to Table 3-18:]

1. THICKNESS OF PROTECTIVE COVERING OF URBAN TELEPHONE CABLES (SET FORTH BY GOSTEKHNIKA USSR No. 327)
2. Diameter of cable over lead sheathing, mm
3. Nominal thickness of protective covering, mm
4. Bedding
5. Armor
6. Outer covering
7. Steel tape
8. Galvanized steel wire
9. Round
10. Flat
11. 50 and higher

The outer covering of bare armored cables consists of a layer of binding compound and chalk solution.

Table 3-19

Толщина защитных покровов кабелей связи с кордельно-бумажной изоляцией (ГОСТ 5008-49) в мм 1

2 Диаметр кабеля поверх свинцовой оболочки, мм	3 Номинальная толщина защитных покровов					
	4 Подушка для броня из		5 Броня из			6 Наружный покров
	7 ленты и плоских проволок	8 круглых проволок	9 стальных лент	10 оцинкованных стальных проволок		
				11 плоских	12 круглых	
13 До 13 13	1,5	—	2×0,3	—	—	1,5
Свыше 13 до 23	1,5	2	2×0,5	1,5	4	1,5
14 » 23 » 27	2	2	2×0,5	1,5	4	2
» 37 » 50	2	2,5	2×0,5	1,7	6	2
» 50	2,5	2,5	2×0,8	1,7	6	2

1. THICKNESS OF PROTECTIVE COVERINGS OF COMMUNICATIONS CABLES WITH PACKTHREAD/PAPER INSULATION (GOST 5008-49) IN mm
2. Diameter of cable over lead sheathing, mm
3. Nominal thickness of protective coverings (Cont'd)

4. Bedding for armor of
5. Armor of
6. Outer covering
7. Tape and flat wire
8. Round wire
9. Steel tape
10. Galvanized steel wire
11. Flat
12. Round
13. To
14. From

Table 3-20

(1) Толщины защитных покровов кабелей для сигнализации и блокировки (ГОСТ 985-47)

(2) Диаметр кабеля поверх свинцовой оболочки, мм	(3) Номинальная радиальная толщина защитных покровов, мм				
	(5) Подушка	(4) Броня из			(10) Наружный покров
		(6) стальных лент	(7) стальных оцинкованных проволок		
			(8) плоских	(9) круглых	
До 13 (11)	1,5	2×0,3	—	—	1,5
Свыше 13 до 23 (12)	1,5	2×0,5	1,5	4	1,5
(12) „ 23 до 37	2,0	2×0,5	1,5	4	2,0

1. THICKNESSES OF PROTECTIVE COVERINGS OF SIGNALING AND HOLDING CABLES (GOST 985-47)
2. Diameter of cable over lead sheathing, mm
3. Nominal radial thickness of protective covering, mm
4. Armor of
5. Bedding
6. Steel tape
7. Galvanized steel wire
8. Flat
9. Round
10. Outer covering
11. To 13
12. From 13 to 23

B. THE ELECTRICAL PARAMETERS OF CABLES AND THEIR CALCULATION

3-4. Resistance of symmetrical circuits.

In its general form, the resistance of a cable circuit consists of the resistance R_0 to direct current and an additional resistance $R_\sim$ governed by the passage of an alternating current through the circuit.

$$R = R_0 + R_\sim . \tag{3-1}$$

The resistance to direct current depends on the material and diameter of the conductor. The current-conducting properties of the material are conventionally expressed in terms of its specific electrical resistance ρ, which characterizes the resistance, expressed in ohms, of a conductor 1 m long with a sectional area of 1 mm^2.

The reciprocal of the specific resistance is called the specific conductance (conductivity) γ; $\gamma = \frac{1}{\rho}$. The values of ρ and therefore of γ are standardized for 20°C.

The resistance of a conductor is determined from the formula

$$R_0 = \rho \frac{l}{q}, \tag{3-2}$$

where l is the length of the conductor in m;

q is the section of the conductor in mm^2.

Or, for 1 km, $R_0 = \rho \frac{l}{q} = \rho \frac{1\,000}{\frac{\pi d^2}{4}} = \rho \frac{4\,000}{\pi d^2}$ [ohms/km].

The specific-resistance values for certain conductor materials are presented in Table 3-21.

Table 3-21

1 **Основные свойства металлов**

2 Наименование материала	3 Удельное сопротивление ρ, $\frac{ом.мм^2}{м}$	4 Удельная проводимость γ, $\frac{ом.м}{мм^2}$	5 Температурный коэффициент α, 1/°C	6 Удельный в g, $г/см^3$
7 Медь	0,0175	57	0,004	8,9
8 Алюминий	0,291	34,36	0,0043	2,65
9 Сталь	0,139	7,23	0,006	7,9

1. BASIC PROPERTIES OF METALS
2. Material
3. Specific resistance ρ, ohms-mm^2/m
4. Specific conductance γ, ohm-m/mm^2
5. Temperature coefficient α, $^\circ C^{-1}$
6. Specific gravity g, g/cm^3
7. Copper
8. Aluminum
9. Steel

When we use the listed values of ρ for copper and aluminum, the formulas for calculation of the d-c resistance of a two-conductor cable circuit takes the form

(for copper conductors)

$$R_0 = 2\rho\frac{4\,000}{\pi d^2} = \frac{44{,}8}{d^2} \quad [\text{ohms/km}];$$

(for aluminum conductors)

$$R_0 = \frac{74}{d^2} \quad [\text{ohms/km}].$$

Copper, which possesses superior electrical con-

ductivity, is used most widely in cable technology.

The use of aluminum involves an increase in the sectional area of the conductors by a factor of approximately 1.65, and this, in turn, increases the bulk of the cable and, consequently, the outlay of lead and other protective armoring materials.

Table 3-22 presents the conductor diameters of copper and aluminum cable circuits having equivalent basic electrical properties.

Table 3-22

CONDUCTOR DIAMETERS OF EQUIVALENT COPPER AND ALUMINUM CABLE CIRCUITS

Diameter of copper Conductors, mm	Corresponding diameter of Aluminum Conductors, mm
0,9	1,15
1,2	1,55
1,4	1,80

Due to the twist, the true length of the conductors is always greater than the length of the cable; this leads to an increase in the resistance of the circuit. The increase in conductor resistance due to the twist is presented in Table 3-23 as standardized by MKK.

The resistance R of the conductors is a function of

temperature, and increases with it.

The resistance of a cable circuit at a temperature

Table 3-23

INCREASE IN CONDUCTOR RESISTANCE
AS A RESULT OF TWISTING

Diameter of layer, mm	Increase, %
До 30	1
30—40	1,5
40—50	2,5
50—60	3,7
60—70	5,0
70—80	7,0

other than 20°C is determined from the formula

$$R_t = R_{20}[1 + \alpha(t - 20)] \quad [\text{ohms/km}], \qquad (3\text{-}3)$$

where R_{20} is the resistance of the circuit at $t = 20°C$;

α is the temperature coefficient;

t is the temperature in question.

Values of the temperature coefficient α are listed in Table 3-21.

The additional resistance of the conductors $R_\sim$ that appears on passage of an alternating current through them results from the creation of alternating electromagnetic fields and the eddy currents to which they give rise in the conductors and in the other metallic parts of the cable surrounding the transmission circuit.

It is demonstrated in the theory of electricity that

when an alternating magnetic field acts on a conducting body, an induction emf arises in this conductor and creates parasitic (eddy) currents.

The eddy currents form closed loops in the interior of the metal about the lines of force of the alternating magnetic field. According to Lenz's law, the direction of these currents is opposed to that of the magnetic field which induces them.

As the magnetic field H rises, the direction of the eddy currents formed about the lines of force is given by the left-hand rule* and these currents create, in turn, a magnetic field $H_{v.t}$ [$H_{eddy-current}$] which is opposed to the basic magnetic field.

As the magnetic field H falls off, the direction of the eddy currents is given by the right-hand rule ["coincides with the motion of the corkscrew handle"] and the magnetic field $H_{v.t}$ which they create tends to support the basic magnetic field.

In our case it is sufficient to consider the action of the eddy currents for the rising field alone, since the declining field is subject to the same laws.

The effect of the eddy currents is proportional to the frequency of the current being transmitted, and also

*"is opposite to the direction of the corkscrew handle."

to the conductivity γ_1 and the magnetic permeability μ_1 of the metal of the conductor.

Quantitatively, these currents are expressed by the equation

$$\sigma = \sqrt{j}K = \sqrt{j\omega\mu_1\gamma_1} = \frac{K}{\sqrt{2}} + j\frac{K}{\sqrt{2}} = |K|\, e^{j45^\circ},$$

where the absolute magnitude $|K|$ of the eddy-current factor characterizes the loss of energy in the interior of the metal, and the 45° angle indicates the phase shift of the current in passage through the metal.

As we know, the eddy currents cause a loss of energy in heating the conductor; this is offset at the expense of the electromagnetic energy transmitted through the circuit. The eddy-current energy loss is determined by the expression

$$W_{v.t} = I^2 R_{\sim}, \qquad [v.t = \text{eddy current}]$$

where I is the current passing through the transmission circuit;

$R_{\sim}$ is the loss resistance, which governs the increase in the resistance of the circuit.

These losses occur as a result of the fact that the eddy currents occasion a redistribution of the electromagnetic field in the interior of the conductor, i.e., a change in the current density over its cross section.

The following three cases are to be distinguished

in accordance with the origin of the eddy currents and their effects:

1. Eddy currents formed in the conductor due to the internal magnetic field of the current passing through it.

Here the lines of force of the internal magnetic field, cutting across the mass of the conductor, induce in it eddy currents which are directed, in accordance with Lenz's law, in opposition to the rotation of the corkscrew handle.

As shown in Fig. 3-14, the eddy currents $I_{v.t}$ in the center of the wire are directed against the basic current flowing in the wire, while their directions coincide at the periphery.

The result of the interaction of the eddy currents with the basic current is a redistribution of the current over the section of the conductor in such a way that the current density increases toward the surface of the wire.

This phenomenon bears the name "skin effect".

The skin effect is directly proportional to the frequency of the current and the magnetic permeability and diameter of the conductor. It is more strongly manifested in steel conductors than in copper conductors. The result of this effect is that at a sufficiently high frequency, the current flows only along the peripheral part of the

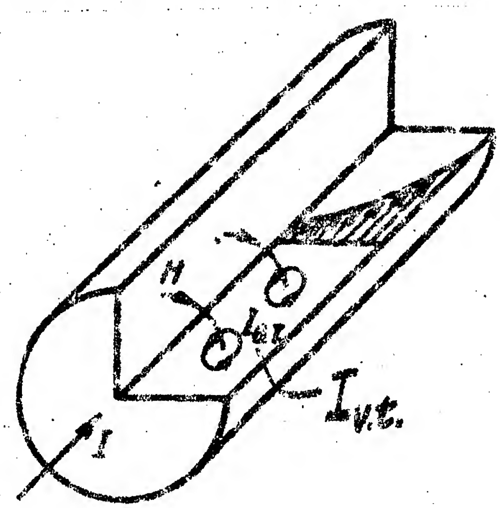


Fig. 3-14. Skin-effect phenomenon.
[$I_{v.t}$ = eddy current]

conductor's section, and this naturally causes an increase in its resistance by an amount $R_{p.e}$ [$R_{skin\ effect}$].

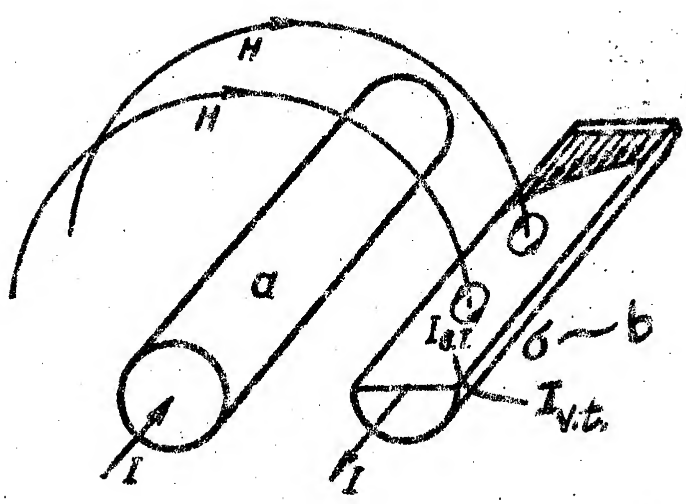


Fig. 3-15. Convergence effect.
[$I_{v.t}$ = eddy current]

2. Eddy currents appearing in one conductor of a two-conductor circuit due to the external magnetic field of the current flowing in the other. As will be seen from Fig. 3-15, the external magnetic field of conductor a, in cutting across the mass of conductor b, induces eddy currents in the latter.

At the surface of conductor b which is turned toward conductor a, the eddy currents coincide in direction with the basic current $(I + I_{v.t})$ flowing through it, while at the averted surface of conductor b they are directed against the basic current $(I - I_{v.t})$. A similar redistribution of currents occurs in conductor a.

As a result of the interaction of the eddy currents with the basic current, the current density at the facing surfaces of conductors a and b increases, and that at the averted surfaces declines. This phenomenon ("convergence" of the currents in the conductors a and b toward one another) bears the name "proximity effect".

Here, just as in the case of the skin effect, only part of the sectional area of the conductors is used as a result of nonuniform distribution of current density, and this increases the resistance of the circuit to alternating current by an amount R_{bl} [$R_{proximity}$].

The action of the proximity effect is also directly proportional to frequency, magnetic permeability, conductivity, and the diameter of the conductor, and, in addition,

is heavily dependent on the distance between the conductors. As the conductors approach one another, the proximity effect increases in a square-law relationship.

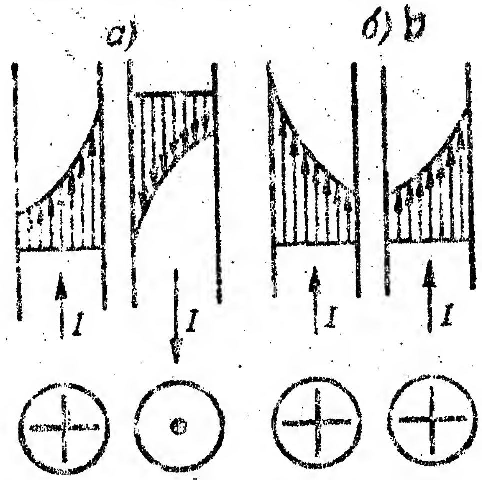


Fig. 3-16. Current-density distribution in pair.
a--symmetrical transmission; b--non-symmetrical transmission.

This accounts for the fact that in cable circuits in which the conductors are placed close together, the proximity effect exerts a strong influence on the quality of the transmission, while in aerial communications lines, where the conductors are considerably more remote from one another, it [the effect] is usually disregarded.

It should be noted that if the currents in two adjacent conductors flow in the same direction, the redistribution of their densities due to the interaction of their external magnetic fields results in a displacement of the currents toward the averted surfaces of the con-

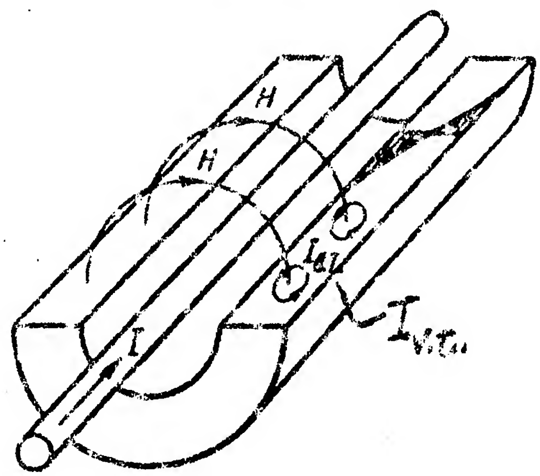


Fig. 3-17. Eddy currents in screen and lead sheath of cable.
[$I_{v.t}$ = eddy current]

ductors a and b.

Fig. 3-16 shows the distribution of current densities in the conductors of a symmetrical circuit when the currents in conductors a and b are opposed and when they flow in the same direction.

The action of the proximity effect conforms to the law of interaction of magnetic poles, according to which like poles repel each other and unlike poles attract each other. This effect also results in drawing together of opposed currents and divergence of currents flowing in the same direction.

3. Eddy currents set up by an external magnetic field in the metallic media surrounding the circuit under study.

The magnetic field created by the current flowing through the circuit induces eddy currents in nearby conductors of the cable, in the surrounding screens, the lead sheath, the armor, etc. (Fig.3-17)

In this case we also have a redistribution of current density over the section of the conductors. Here, however, since a whole series of factors are in simultaneous operation, the pattern of current-density distribution becomes extremely complex.

The eddy currents heat the metallic components of the cable and cause significant additional thermal energy losses; this is expressed, so to speak, in a "siphoning off" of a certain fraction of the transmitted energy, with only the metallic parts of the cable located close to the circuit in question having essential significance.

These losses are also taken into account as an additional resistance R_m of the cable circuit.

Thus when alternating currents are transmitted through a cable circuit, its total resistance will be composed as follows:

$$R = R_0 + R_{\sim} = R_0 + R_{\text{п.э}} + R_{\text{бл}} + R_{\text{м}}.$$

[last 3 terms: R_{skin}, R_{prox}, R_{metal}].

To compute the resistance of the circuit taking the skin and proximity effects in account, the following for-

mula should be used:

$$R = 2R_0\left[1 + F(Kr) + \frac{pG(Kr)\left(\frac{d}{a}\right)^2}{1 - H(Kr)\left(\frac{d}{a}\right)^2}\right], \qquad (3\text{-}4)$$

where $2R_0$ is the resistance of the circuit to direct current (the 2 indicates that the resistances of both conductors of the circuit are being considered);

$2R_0 F(Kr) = R_{p.e}$ [R_{skin}] is the additional resistance of the circuit due to the skin effect;

$2R_0 \dfrac{pG(kr)\left(\frac{d}{a}\right)^2}{1 - H(Kr)\left(\frac{d}{a}\right)^2} = R_{bl}$ [R_{prox}] is the additional resistance of the circuit due to the proximity effect between the conductors;

α is the distance between the centers of the conductors in cm;

d = 2r is the diameter of the conductor in cm;

p is a factor taking into account the type of twist.

Values of p for the different twists are given in Fig. 3-18.

The functions F, G, and H of the eddy-current loss factor $K = \sqrt{\omega\mu_1\gamma_1}$ and the radius r of the conductor are presented in Table 3-24.

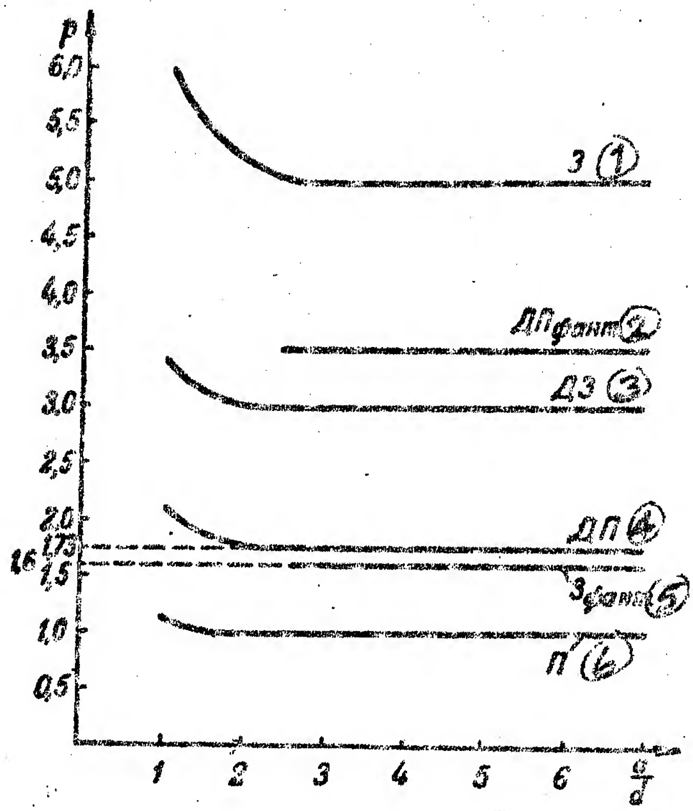


Fig. 3-18. Proximity-effect losses for different types of cable twist.

1--quad; 2--double pair (actual); 3--double quad; 4--double pair; 5--quad (actual); 6--pair

Eddy-current loss factors and values of Kr for conductors of various types are assembled in Table 3-25.

It follows from Table 3-24 that at large values of Kr (in the high-frequency region), G becomes equal to half of F, and H approaches a constant value of 3/4.

Table 3-24

(1) Функции F, G и H для различных значений Kr

Kr	$F(Kr)$	$G(Kr)$	$H(Kr)$	$Q(Kr)$
0	0	$\frac{(Kr)^4}{64}$	0,0417	1
0,5	0,000326	0,000975	0,042	0,9998
1,0	0,00519	0,01519	0,053	0,997
1,5	0,0258	0,0691	0,092	0,987
2,0	0,0782	0,1724	0,169	0,961
2,5	0,1756	0,295	0,263	0,913
3,0	0,318	0,405	0,348	0,845
3,5	0,492	0,499	0,416	0,766
4,0	0,678	0,584	0,466	0,686
4,5	0,862	0,669	0,503	0,616
5,0	1,042	0,755	0,530	0,556
7,0	1,743	1,109	0,596	0,400
10,0	2,799	1,641	0,643	0,282
$>10,0$	$\frac{\sqrt{2}(Kr)-3}{4}$	$\frac{\sqrt{2}(Kr)-1}{8}$	0,750	$\frac{2\sqrt{2}}{Kr}$

1. THE FUNCTIONS F, G, AND H FOR DIFFERENT VALUES OF Kr

Table 3-25

	(1) Медь	(2) Сталь	(3) Алюминий
$K=\sqrt{\omega\mu_1\gamma_1}$	$0,21\sqrt{f}$	$0,75\sqrt{f}$	$0,164\sqrt{f}$
Kr	$0,0105d\sqrt{f}$	$0,0375d\sqrt{f}$	$0,082d\sqrt{f}$

(4) Примечание. В табл. 3-25 размерность диаметра проводника дана в миллиметрах

1. Copper
2. Steel
3. Aluminum
4. Note: the conductor diameters in Table 3-25 are given in mm.

The resistance R_{bl} [R_{prox}] as a function of the distance between the conductors at f = 60,000 cps is shown in Fig. 3 19, whence it follows that when the distance a between the conductors $\geqslant$ 4d, the proximity effect is very weakly

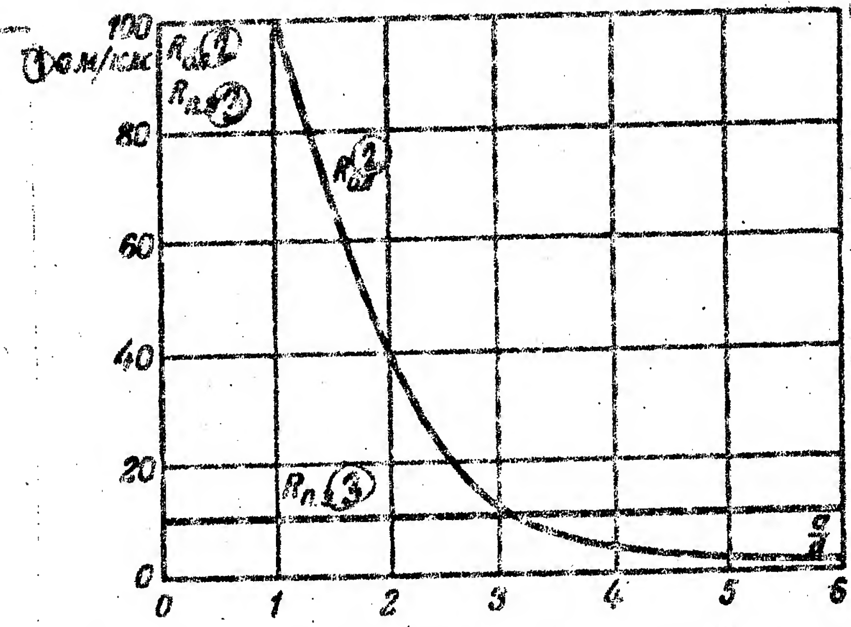


Fig. 3-19. Resistances introduced by proximity and skin effects with various distances between conductors.

1--ohms/km; 2--R_{bl} (R_{prox}); 3--$R_{p.e}$ (R_{skin}).

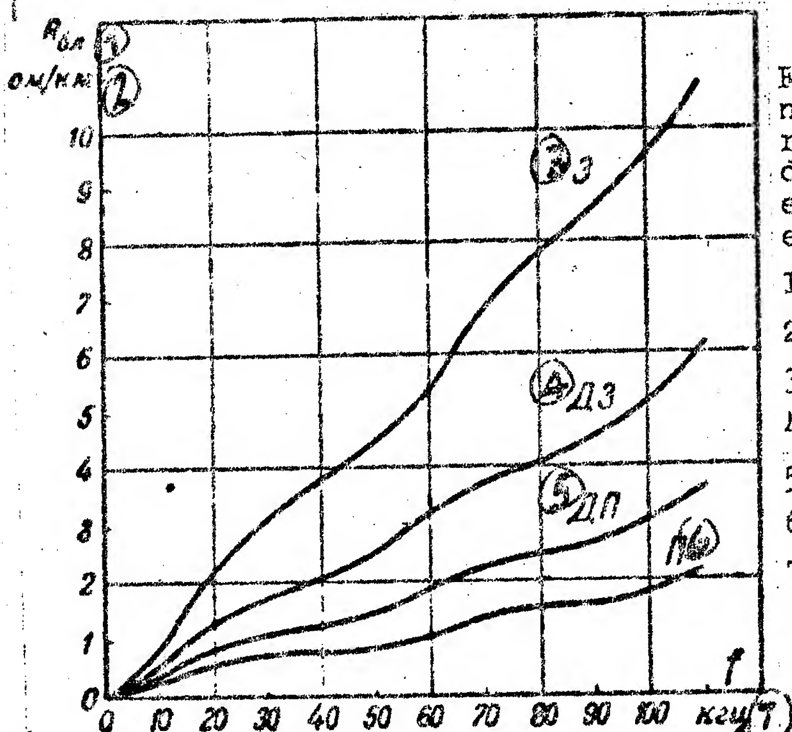


Fig. 3-20. Frequency-dependence of resistance introduced by proximity effect for different types of twist.

1--R_{bl} (R_{prox})

2--ohms/km

3--quad

4--double quad

5--double pair

6--pair

7--kcps

manifested. The resistance $R_{p.e}$ [skin-effect] of a cable circuit with conductors 1.2mm in diameter is given in the same figure for comparison.

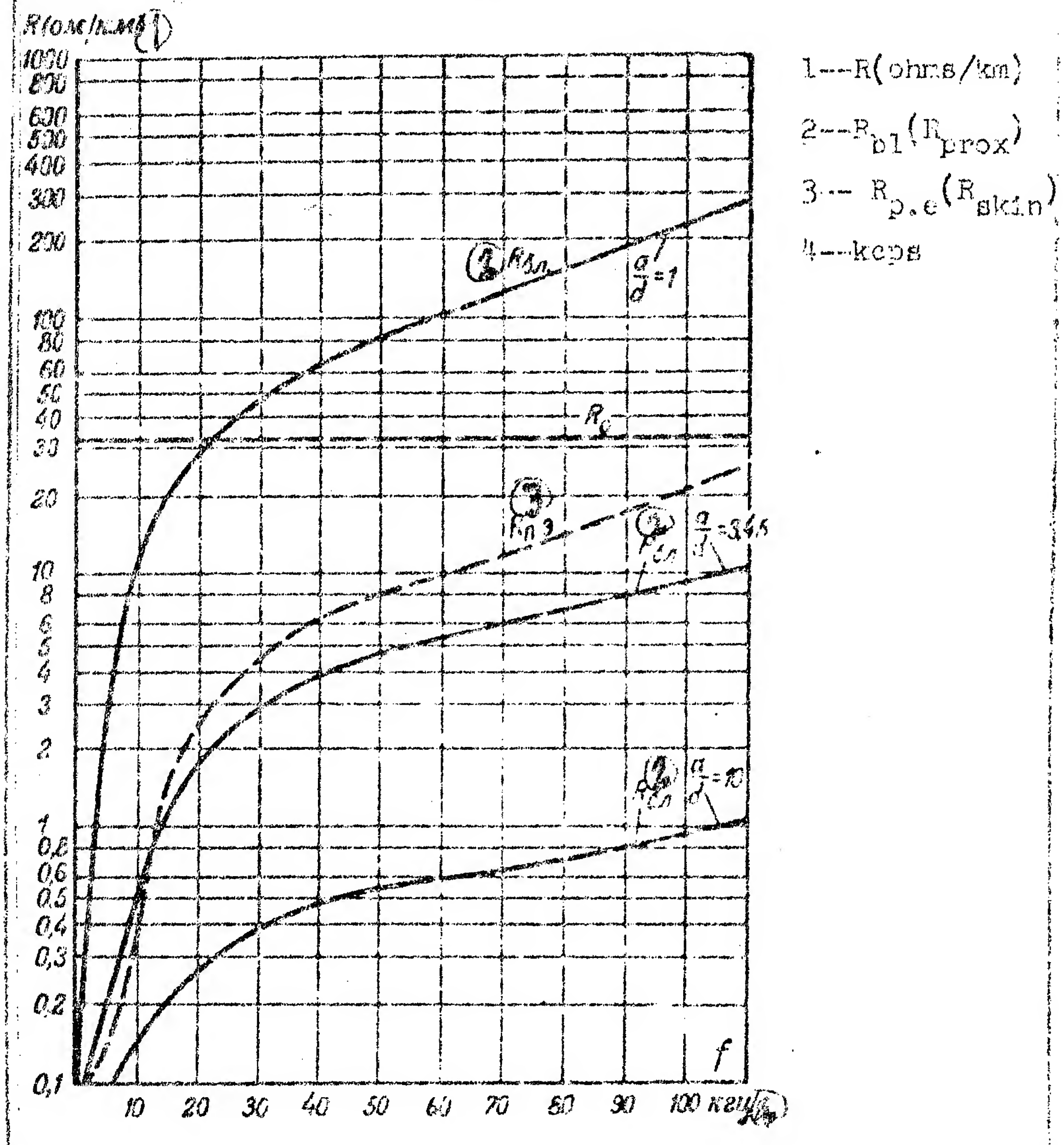


Fig. 3-21 Frequency-dependence of resistances introduced by proximity and skin effects for various distances between conductors (quadded cable with d= 1.2mm).

Type P is most favorable from the standpoint of the proximity effect; it is followed by DP, DZ, and finally Z (Fig. 3-20). The value of R_{bl} is 5 times larger in the spiral quad (Z) than in the pair (P).

Fig. 3-21 shows $R_{p.e}$ and R_{bl} as functions of frequency for a quadded cable with conductors having d = 1.2mm. The value of R_{bl} is calculated for three different distances between conductors ($\frac{a}{d}$ = 1, 3.45, and 10).

The variation of the resistance as the conductor thickness increases from 1.0 to 1.6 mm at f = 60,000 cps is shown in Fig. 3-22. It is seen from this figure that as the conductor diameter increases, the d-c resistance R_0 of the circuit drops sharply, but that a simultaneous increase in the resistance $R_\sim$ takes place due to the eddy-current losses.

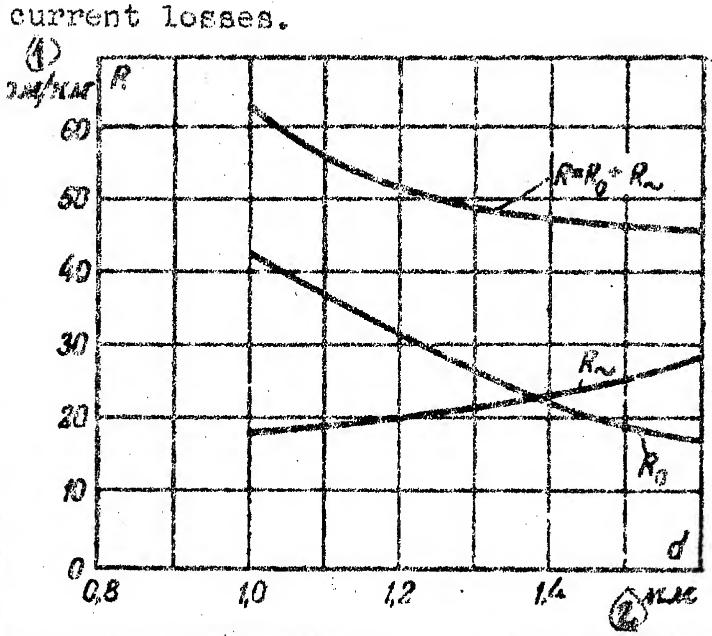


Fig. 3-22. Resistances of cables to direct and alternating current for various conductor diameters.

1--ohms/km

2--mm

Thus when direct current or low-frequency current is transmitted through cables, enlargement of the conductors dives a significant improvement in the sense of increasing the range of communication. In the case of high-frequency communications, however, eddy-current losses sharply reduce the effectiveness of an increase in cable-conductor diameter, and with a given sectional area the gain in range and quality of communication does not justify the outlay of materials involved in enlarging the conductors.

This circumstance is one of the reasons why conductors no larger than 1.2--1.4mm in diameter are used in contemporary long-range communications cables (conductors 2-3mm in diameter had been used previously.)

The increase in the resistance of the conductors introduced by neighboring groups, the lead sheathing, and other metallic parts of the cable into the transmission circuit does not admit of exact calculation, since in this case we have a series of differently-directed magnetic-field components acting simultaneously, and this results in extreme complication of the field's configuration. Therefore it is customary to determine the additional resistance R_m governed by losses in the surrounding metallic parts of the cable by experimental means.

Table 3-26 gives the additional resistances R_m introduced by neighboring groups and the lead sheathing for various cable designs at a frequency of 200,000 cps.

Table 3-26

1 **Дополнительное сопротивление $R_м$ ом/км из-за потерь в токопроводящих жилах смежных четверок и свинцовой оболочке кабеля**

2 Число четверок в кабеле	3 Основная цепь				4 Фантомная цепь			
	1 повив 5	2 повив 6	3 повив 7	4 повив 8	1 повив 5	2 повив 6	3 повив 7	4 пов... 8
9 **а) Сопротивление потерь в смежных четверках**								
1	0	—	—	—	0	—	—	—
1+6	8	7,5	—	—	1,2	1,2	—	—
1+6+12	8	7,5	7,5	—	1,2	1,2	1,2	—
1+6+12+18	8	7,5	7,5	7,5	1,2	1,2	1,2	1,2
10 **б) Сопротивление потерь в свинцовой оболочке**								
1	22	—	—	—	5,7	—	—	—
1+6	1,5	5,5	—	—	0,5	1,7	—	—
1+6+12	0	0	1,0	—	0	0	0,7	—
1+6+12+18	0	0	0	1,0	0	0	0	0,5

1. ADDITIONAL RESISTANCE Rm [ohms/km] DUE TO LOSSES IN THE CONDUCTORS OF ADJACENT QUADS AND IN THE LEAD SHEATHING OF THE CABLE.
2. Number of quads in cable
3. Basic circuit
4. Phantom circuit
5. 1st layer
6. 2nd "
7. 3rd "
8. 4th "
9. a) Loss resistance in adjacent quads
10. Loss resistance in lead sheathing

It follows from the table that, for the most part, the lead sheathing introduces its losses into the circuits of the outermost layer, which is in immediate proximity to it.

The additional loss resistance is significantly higher in the phantom circuits than in the basic circuits.

The value of R_m for other frequencies f is determined from the following empirical formula:

$$R_m = R'_m \sqrt{\frac{f}{200\,000}}, \qquad (3\text{-}5)$$

where R'_m is the tabulated value of the additional resistance due to losses in the adjacent quads or in the lead sheathing at f = 200 cps.

Table 3-27 lists values of the resistance of a 32x2 quadded cable with d = 1.2 mm and a = 4.14 mm as a function of frequency over a wide frequency range. The same table shows the relative values of the various components $R_{p.e}$, R_{bl}, and R_m in the total resistance of the cable.

It is clear from the table that at f = 60,000 cycles, the skin effect increases the resistance of the circuit by 31%, and the proximity effect raises it by 17%. On the whole, the resistance of the cable circuit to alternating current is 1.62 times its resistance to direct

current.

At the frequency f = 108,000 cps, the resistance to alternating current is somewhat more than twice as large as the resistance to direct current.

Table 3-27

(1) **Активное сопротивление кабеля звездной скрутки с диаметром жил 1,2 *мм***

(2) f, *кгц*	(3) R_0, *ом/км*	(4) $R_{общ}$, *ом/км*	(5) $R_{п.э}$, *ом/км*	(6) $R_{бл}$, *ом/км*	(7) $R_м$, *ом/км*	(8) R_0 в % от $R_{общ}$	(9) $R_{п.э}$ в % от $R_{общ}$	(10) $R_{бл}$ в % от $R_{общ}$	(11) $R_м$ в % от $R_{общ}$
0,8	31,6	31,79	0,0083	0,0101	0,17	99,45	0,026	0,023	0,6
5	31,6	33,25	0,133	0,161	1,36	95,2	0,4	0,48	4,1
13,5	31,6	35,53	0,815	0,908	2,21	89	2,2	2,5	6,3
20	31,6	37,7	1,81	1,735	2,55	83,83	4,8	4,6	6,77
30	31,6	41,57	3,7	2,92	3,32	76,2	8,9	7,1	7,98
40	31,6	44,91	5,55	3,95	3,81	70,4	12,3	8,8	8,5
50	31,6	48,89	8,2	4,84	4,25	64,7	16,7	9,9	8,7
60	31,6	51,6	10	5,44	4,6	61,2	19,4	10,5	8,9
70	31,6	56,37	13,6	6,14	5,03	55	24,15	10,9	8,95
80	31,6	50,93	16,2	6,76	5,37	52,8	27	11,14	8,97
90	31,6	63,3	18,7	7,3	5,7	50	29,5	11,5	9
100	31,6	66,7	21,2	7,9	6,0	47,4	31,8	11,8	9
108	31,6	68,85	22,9	8,2	6,15	46,0	33,2	11,9	8,9

1. RESISTANCE OF QUAD CABLE WITH 1.2-mm CONDUCTOR DIAMETER
2. f, kilocycles
3. R_0, ohms/km
4. R_{total}, ohms/km
5. $R_{p.e}$, ohms/km [skin effect]
6. R_{bl}, ohms/km [proximity effect]
7. R_m, ohms/km [metal losses]
8. R_0 as percentage of R_{total}
9. $R_{p.e}$ as percentage of R_{total}
10. R_{bl} as percentage of R_{total}
11. R_m as percentage of R_{total}

3-5. THE INDUCTANCE OF SYMMETRICAL CIRCUITS

According to the law of electromagnetic induction, any change in the magnetic flux linked with any circuit induces a voltage in the latter. Here it is of no consequence whether the magnetic flux is set up by a current in some nearby circuit or in the subject circuit itself.

In the former case, when the induced voltage is created by the alternating magnetic flux of a neighboring circuit, this effect bears the name mutual induction.

The mutual-induction coefficient is expressed by the relationship

$$M=\frac{\Phi_2}{I_1},$$

where Φ_2 is the magnetic flux created by the current flowing in the adjacent circuit, and I_1 is the current in the circuit under consideration.

Self-induction is the phenomenon of induced voltage due to the variation of the conductor's own magnetic field. Self-induction may be represented physically as follows: The alternating current flowing through the circuit creates a magnetic flux of changing direction. In turn, the magnetic lines of force of this flux cross the circuit of the current which has created them and give rise

to the induced-voltage effect in this circuit.

Numerically, the self-inductance coefficient (or inductance) is expressed as the ratio of the magnetic flux to the current creating it:

$$L = \frac{\Phi_1}{I_1}.$$

The inductance of a cable line is quite analogous in nature to the inductance of a solenoid, but instead of many small turns we deal here with only a single large turn (the cable circuit).

The inductance L is characterized by the shape, dimensions, material, and arrangement of the conductors and is measured as the flux of magnetic induction flowing across a contour bounded by the conductors of the circuit.

The circuit inductance consists of two parts, the external inductance and the internal inductance:

$$L = L_1 + L_2.$$

The external inductance L_1 is governed by the external magnetic field (that outside the conductors) and is determined as the ratio of the external magnetic flux to the current flowing in the circuit.

The internal inductance L_2 is defined as the ratio of the internal magnetic flux (the flux in the interior of

the conductors themselves) to the current flowing in the circuit.

The inductance of two-conductor cable circuits is computed by the formula

$$L = L_1 + L_2 = \left[4 \ln \frac{2a - d}{d} + Q(Kr) \right] \cdot 10^{-4} \text{[henries/km]}, \quad (3\text{-}6)$$

where a is the distance between the centers of the conductors;

d is the diameter of the conductor.

Q(Kr) is a function that depends on the eddy-current factor $K = \sqrt{\omega \mu_1 \gamma_1}$ and the radius r of the conductor.

Values of Q(Kr) are listed in Table 3-24.

It is seen from the formula that the external inductance depends on the radius of the conductors and the distance between them.

The internal inductance is governed by the properties of the cable conductor itself (r, γ_1, μ_1) and the frequency of the transmitted current.

Table 3-28 shows inductance as a function of frequency.

As the frequency of the transmitted current is increased, the total inductance of the circuit drops as a result of the drop in internal inductance. This is accounted for by the fact that as the frequency rises, the

skin effect draws the current to the conductors' surfaces. The magnetic flux is redistributed accordingly (there is none in the center of the wire) and this results in a reduction in the internal inductance of the circuit.

Analytically, the frequency-dependence of inductance is governed by the variation of the function Q(Kr), which, as follows from Table 3-24, is unity for direct current and tends to zero with increasing frequency.

Table 3-28

(1) Частотная зависимость индуктивности кабеля, имеющего $d = 1,2$ *мм* и $a = 4,14$ *мм*

(2) f, *кгц*	Kr	$Q(Kr)$	(3) $L_{внутр}$, *мгн/км*	(4) $L_{внешн}$, *мгн/км*	(5) $L_{общ}$, *мгн/км*
0,8	0,356	1,0	0,102	0,722	0,824
5	0,89	0,998	0,102	0,722	0,824
13,5	1,46	0,997	0,100	0,722	0,822
20	1.78	0,972	0,099	0,722	0,821
30	2,18	0,942	0,096	0,722	0,818
40	2,52	0,913	0,093	0,722	0,815
50	2,82	0,874	0,089	0,722	0,811
60	3,08	0,829	0,085	0,722	0,807
70	3,33	0,794	0,081	0,722	0,803
80	3,56	0,752	0,077	0,722	0,799
90	3,78	0,72	0,073	0,722	0,795
100	3,99	0,688	0,070	0,722	0,792
108	4,14	0,664	0,068	0,722	0,790

1. INDUCTANCE OF CABLE WITH d= 1.2 mm AND a = 4.14 mm AS A FUNCTION OF FREQUENCY
2. f, kilocycles

(Key Cont'd next page)

3. L_{int}, millihenries/km
4. L_{ext}, millihenries/km
5. L_{total}, millihenries/km

External inductance is not a function of frequency.

The result is that with increasing frequency, the total inductance of the circuit drops to a value equal to the external inductance.

Below we consider inductance values of cable circuits for various diameters of the conductors and distances between them.

As the distance a between the conductors increases, the external inductance increases as $\ln \frac{2a - d}{d}$ (Fig. 3-23). This is accounted for by the fact that the area $S = la$ encompassed by the magnetic lines of force increases with increasing a, so that the external magnetic flux Φ increases accordingly:

$$\Phi = BS = Bla. \tag{3-7}$$

As will be seen from Fig. 3-24, the larger a, the larger will be the area penetrated by the magnetic field, and the greater the extent to which the lines of force encompass the circuit. It is for this reason that the inductance of aerial communications lines is about 3 times that of cable lines, and amounts to 2-3 millihenries

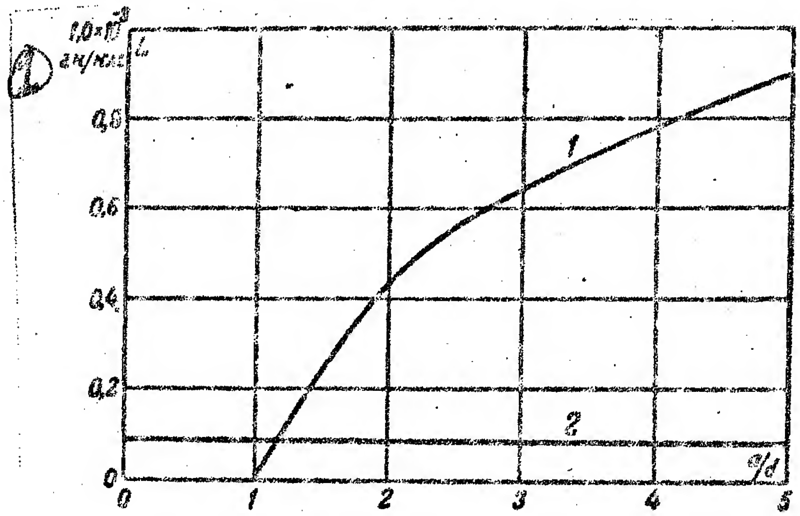


Fig. 3-23. Cable inductance as a function of distance between conductors.

1--external inductance; 2--internal inductance.

Key: 1 = henries per kilometer

per kilometer.

Inductance values for cable circuits with conductors of various diameters are given in Fig. 3-25. It is seen from the diagram that the inductance of the circuit declines somewhat as the cable-conductor diameter increases. This is the result of the fact that first, the skin effect is more strongly manifested with large diameters, so that the internal inductance falls off, and secondly, the area penetrated by the external magnetic flux decreases with increasing conductor diameter, so that the external inductance decreases accordingly.

The inductance of steel circuits is considerably

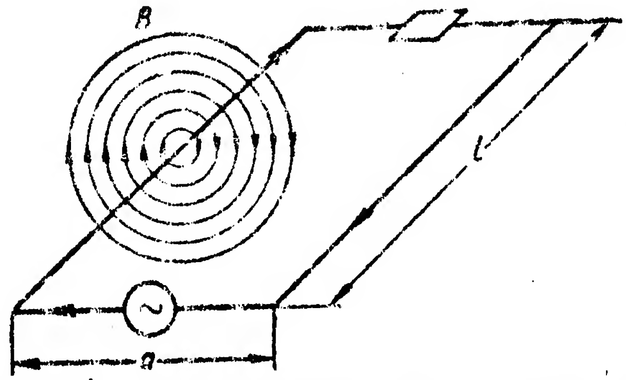


Fig. 3-24. Illustrating calculation of external inductance of cable circuit.

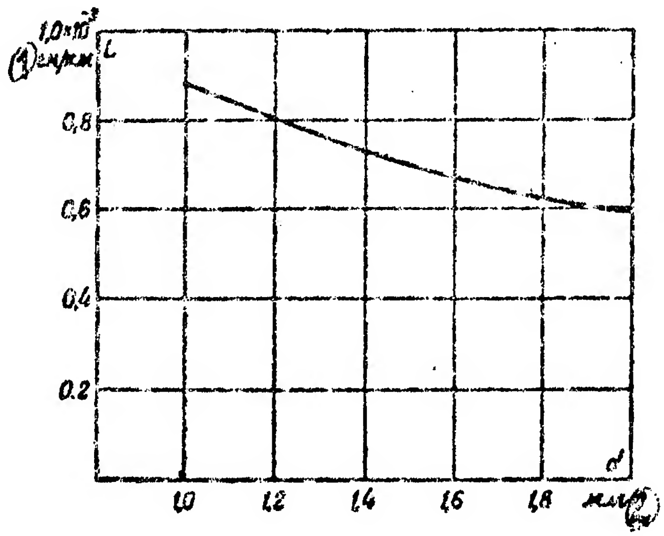


Fig. 3-25. Inductances of cables with conductors of various diameters.

1--henries/km; 2--mm

higher magnetic permeability μ_1. The greater μ_1, the higher will be the concentration of magnetic lines of force within the conductor and the internal inductance of the circuit.

In alternating-current circuits, in which rapidly

varying magnetic fields are created, self-induction exerts a continuous influence on the operation of the circuit.

The latter plays the same role in electrical processes that inertia plays in mechanical systems. By Lenz's law, the emf arising as a result of self-induction is always so directed that it opposes the factor giving rise to it. The electromotive force of self-induction and the emf of the basic current flowing through the circuit are opposed in direction.

It follows from this that self-induction is, as it were, an additional component of resistance to the passage of alternating current. The higher the frequency, and therefore the rate of variation of the magnetic flux, the higher will be the inductive reactance of the circuit:

$$X_L = \omega L.$$

Thus the impedance of the alternating-current circuit, taking the ["active"] resistance R into account, is expressed in the form

$$Z = R + j\omega L.$$

For f = 0 (direct current), the inductive reactance disappears and the impedance ["full resistance"] of the circuit is simply Z = R.

3-6. CAPACITANCE OF SYMMETRICAL CIRCUITS

The capacitance of a cable is analogous to the capacitance of a condenser, with the surfaces of the conductors performing the function of the plates and the insulation between them (air, paper, styroflex, polyethylene, etc.) serving as the dielectric.

Capacitance is the proportionality coefficient between the quantity of electricity Q and the voltage U and characterizes the amount by which the voltage between the capacitor's plates increases when a certain charge

$$C = \frac{Q}{U}.$$

is imparted to it.

The capacitance of a condenser increases as the quantity of electricity accumulated in it at a definite potential difference.

The capacitance of communications cables is measured and standardized in microfarads (10^{-6} fd), nanofarads (10^{-9} fd), and micromicrofarads (10^{-12} fd).

Cable technique distinguishes between two forms of capacitance: a) effective capacitance, i.e., the capacitance between the conductors of the circuit (pair) under consideration, and b) the direct capacitance between the individual conductors of the cable.

The direct capacitances are components of the effective capacitance. In the cable quad shown in Fig. 3-26, the resultant capacitances between the conductors 1-2 and 3-4 are the effective capacitances C_I and C_{II}. The capacitances between the separate conductors---C_{13}, C_{14}, C_{23}, C_{24}, C_{12}, and C_{34}-- and with respect to the lead sheathing of the cable (to ground)--C_{10}, C_{20}, C_{30}, and C_{40} are direct capacitances.

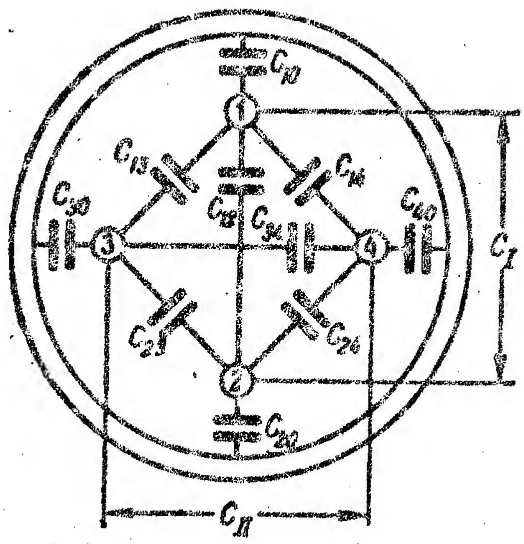


Fig. 3-26. Effective and direct capacitances of cable quad.

The number of direct capacitances, N, may be computed by the formula

$$N = \frac{n}{2}(n+1), \qquad (3\text{-}8)$$

where n is the number of conductors in the cable.

Thus a cable pair will contain three direct capacitances and a quad ten.

The effective capacitance is the fundamental (standardized) quantity determining the quality of transmission through the cable. As shown by V. N. Kulechov, it is calculated by the formula.

$$C = \frac{\lambda\varepsilon}{36 \ln \frac{2a}{d}\psi} 10^{-6} \quad [\text{fd/km}], \qquad (3\text{-}9)$$

where a is the distance between the centers of the conductors (taken in accordence with Table 3-29);

d is the diameter of the conductor;

ψ is a correcting factor characterizing the distance of the conductor from the grounded sheathing (at great distances, $\psi = 1$);

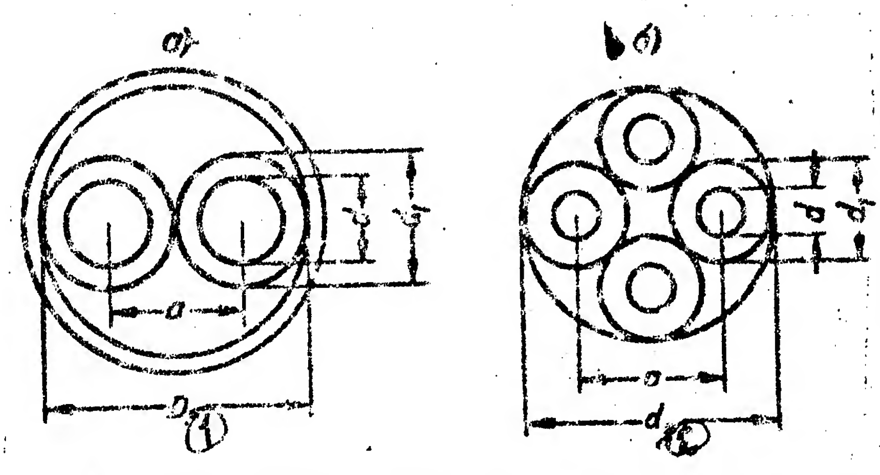


Fig. 3-27. Calculation of cable capacitance.
a--screened pair; b--spiral quad
1--D_e (screen diameter); 2--d_z (quad diameter)

λ is the spiraling factor

ε is the dielectric constant.

Values of a and ψ for various types of cable twist are listed in Table 3-29.

Table 3-29

(1) Значения коэффициентов a и ψ для расчета емкости

№ пор. (2)	Тип скрутки (3)	a	ψ
1	Пара (4)	d_1	$\psi_{П} = \frac{(d_{П} + d_1 - d)^2 - a^2}{(d_{П} + d_1 - d)^2 + a^2}$
2	Звезда (5)	$d - d_1$	$\psi_{З} = \frac{(d_{З} + d_1 - d)^2 - a^2}{(d_{З} + d_1 - d)^2 + a^2}$
3	Четверка ДП (6)	d_1	$\psi_{ДП} = \frac{(0,65 d_{ДП} + d_1 - d)^2 - a^2}{(0,65 d_{ДП} + d_1 - d)^2 + a^2}$
4	Восьмерка ДЗ (7)	d_1	$\psi_{ДЗ} = \frac{(0,43 d_{ДЗ} + d_1 - d)^2 - a^2}{(0,43 d_{ДЗ} + d_1 - d)^2 + a^2}$
5	Экранированная пара (8)	d_1	$\psi_{ПЭ} = \frac{D_э^2 - a^2}{D_э^2 + a^2}$
6	Экранированная звезда (9)	$\sqrt{2}\, d_1$	$\psi_{ЗЭ} = \frac{D_э^2 - a^2}{D_э^2 + a^2}$

1. VALUES OF COEFFICIENTS a AND ψ FOR CALCULATION OF CAPACITANCE
2. No.
3. Type of twist
4. Pair П
5. Spiral four З
6. Double-pair four ДП
7. Octuplet ДЗ
8. Screened pair ПЭ
9. Screened Spiral four ЗЭ

where $d_{П}$, $d_{З}$, $d_{ДП}$, and $d_{ДЗ}$ are the diameters of the groups in the corresponding twist systems;

d_1 is the diameter of the insulated conductor;

a is the distance between the conductor centers

d is the diameter of the conductors;

D_e is the diameter of the screen.

Fig. 3-27 shows these designations for the spiral quad and the screened pair.

Table 3-30 gives numerical values of the coefficient φ for various types of twist.

Table 3-30

VALUES OF COEFFICIENTS φ FOR VARIOUS RATIOS $\frac{a}{d_1}$

$\frac{a}{d_1}$	$\varphi_{П}$	$\varphi_{З}$	$\varphi_{ДП}$
1,6	0,608	0,588	0,615
1,8	0,627	0,611	0,528
2,0	0,644	0,619	0,660
2,2	0,655	0,630	0,670
2,4	0,665	0,647	0,692

П = pair
З = quad
ДП = double pair

In addition to formula (3-9), the following empirical formula is widely applied for the calculation of effec-

tive capacitance:

$$C = \frac{\varepsilon}{36 \ln \frac{aD}{d}} 10^{-6} [\phi/\kappa м],$$

$$[fd/km], \quad (3\text{-}10)$$

where D is the diameter of the group;

d is the diameter of the conductor;

α is a correcting factor which depends on the system in which the conductors are twisted into the group.

The value of α is

0.94 for the pair;

0.75 for the quad;

0.65 for the double pair.

The result of calculation by Formula (3-10) agrees sufficiently closely with the actual values only for groups inside a strand of conductors. For groups in the outermost layer next to the lead sheathing or in proximity to screened groups, the calculation gives results somewhat on the low side.

It is interesting to trace the dependence of capacitance on the variation of the conductor diameter d and the interconductor distance a.

Table 3-31 gives theoretical capacitance values for cables with packthread-paper insulation and conductors of different diameters (from 1.0 to 1.6 mm). In the calculation, the group diameter d_z $[d_{quad}]$ is assumed to be constant at 7.07. The value of ε is 1.38.

Table 3-31

VARIATION OF CAPACITANCE WITH INCREASING
CABLE-CONDUCTOR DIAMETER

(1) d, мм	1,0	1,2	1,4	1,6
$\frac{a}{d}$	3,45	3,44	3	2,85
(2) C, нф/км	23,3	26,5	29,8	32

1--d, mm 2--C, nanofarads/km

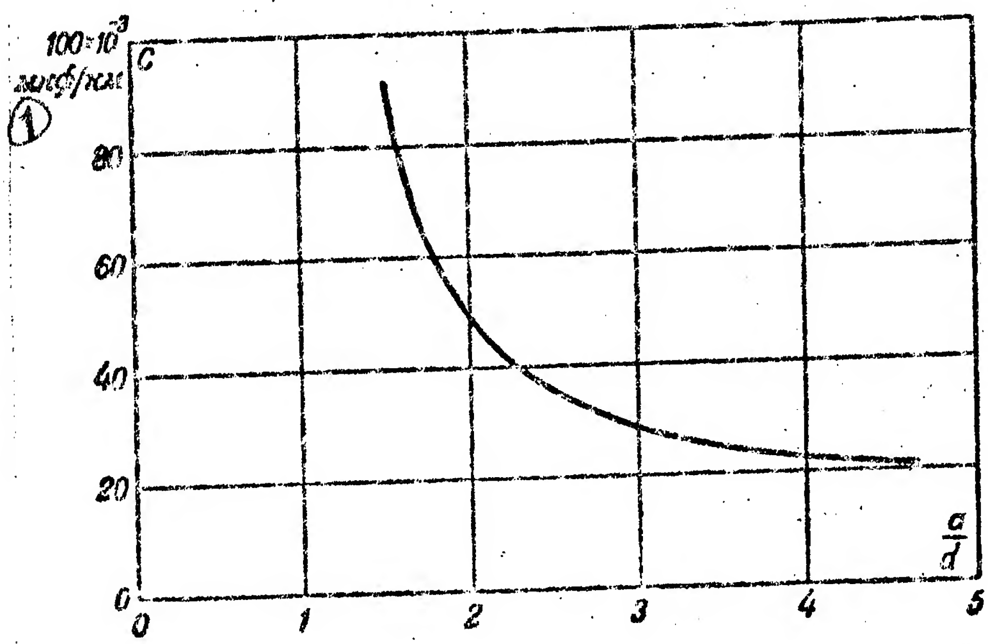


Fig. 3-28. Variation of capacitance with increasing distance between cable conductors.
1--microfarads/km

It follows from the table that the capacitance of the cable increases with increasing conductor diameter. This is consistent with the physical sense of the phenomenon under consideration, since an increase in conductor

diameter is equivalent to enlargement of the plates of a condenser, i.e., results in increased capacitance.

It follows from Fig. 3-28, which shows the variation of a cable's capacitance with increasing distance between the conductors, that as the distance a between the conductors increases, the capacitance of the cable decreases considerably. Actually, an increase in the distance a in cables is equivalent to moving the plates of a capacitor farther apart, and this naturally gives a reduction in capacitance.

The parameter ϵ appearing in the formulas for calculating capacitance is the resultant value of the dielectric constant for the composite dielectric used in the cable.

3-7. SHUNT CONDUCTANCE (LEAKANCE) OF SYMMETRICAL CIRCUITS

The shunt conductance G is an electrical parameter of the line which characterizes the quality of the insulation covering the cable's conductors.

Just as the resistance determines the loss of electromagnetic energy in the metallic parts of the cable, the parameter G indicates the loss of energy in the insulation of its conductors.

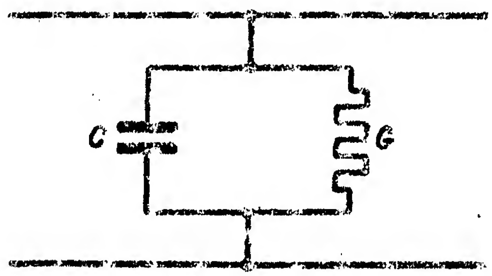


Fig. 3-29. Illustrating calculation of shunt conductance.

The shunt conductance of a line may be represented as a resistance equivalent that shunts its capacitance (Fig. 3-29).

The resistance of the insulation and its conductance are related by the expression $G = 1/R_{ins}$.

The shunt conductance of a line is governed by the properties of the cable dielectrics and primarily by the resistivity ρ and the dielectric loss factor tan δ. Since the insulation of the conductors of cables possesses a certain electrical conductivity, part of the current does not reach the end of the line, completing its circuit between the conductors and being dissipated in the dielectric (the so-called "leakage" of current).

Direct- (G_0) and alternating- ($G_\sim$) current shunt conductances are distinquished in accordance with the nature and origin of the effect. The shunt admittance is expressed as the sum $G = G_0 + G_\sim$.

The quantity G_0 is inversely proportional to the direct-current insulation resistance of the cable circuit:

$$G_0 = \frac{1}{R_{ins}}.$$

The presence of the conductance G_0 is accounted for by the nonideal electrical properties of the cable dielectric. Their volume and surface resistivities have finite values ($\rho = 10^{12}$ to 10^{17} ohms-cm), with the result that electric charges can migrate through the dielectric and a path is created for leakage of direct current between the forward and return conductors of the cable circuit and to ground. This conductance causes energy losses in heating the dielectric which are determined by the expression

$$P_0 = U^2 G_0 = \frac{U^2}{R_{ins}}. \qquad (3\text{-}11)$$

The energy losses P_0 in the dielectric due to direct-current shunt conductance are analogous in principle to the thermal energy loss in the cable conductors.

The phenomenon of dielectric polarization consists in the formation of dipoles which change direction (shift) during the cycle of variation of the applied electromagnetic field.

The dipole-shifting process results in a loss of energy expressed by the formula $P_{\sim} = U^2 G_{\sim}$. Under certain

conditions, $P_{\sim}$ may attain values exceeding the energy loss in the cable conductors.

It is customary to express the value of $G_{\sim}$ and, accordingly, the polarization losses in solid dielectrics in terms of the dielectric loss factor $\tan\delta$, which is determined experimentally. The former is directly proportional to the frequency of the transmitted current, the capacitance of the cable, and the dielectric loss factor:

$$G_{\sim} = \omega C \tan\delta.$$

The dielectric loss constant is the most important parameter in determining the possibility of using a dielectric in a communications cable.

For high-quality dielectrics with insignificant loss constants, it may be assumed that $\tan\delta \approx \delta$, since the tangent is equal to its argument for small angles.

Values of the dielectric loss constant for various cable dielectrics are listed in Chapter 9.

The formula for the full shunt conductance takes the form

$$G = G_0 + G_{\sim} = \frac{1}{R_{ins}} + \omega C \tan\delta. \quad (3\text{-}12)$$

The shunt conductance is measured in mhos/km (reciprocal resistance).

In comparing the values of G_0 and $G_{\sim}$, it should be

noted that dielectric-polarization losses in communications cables are significantly larger than thermal losses due to nonideal insulation. Therefore, the conductance G_0 may be disregarded and the calculation performed by the formula

tan

$$G = G_{\sim} = \omega C\,\mathrm{tg}\,\delta. \qquad (3\text{-}13)$$

It is necessary to consider the quantity G_0 only when considering the cable transmission of direct current and telegraph signals.

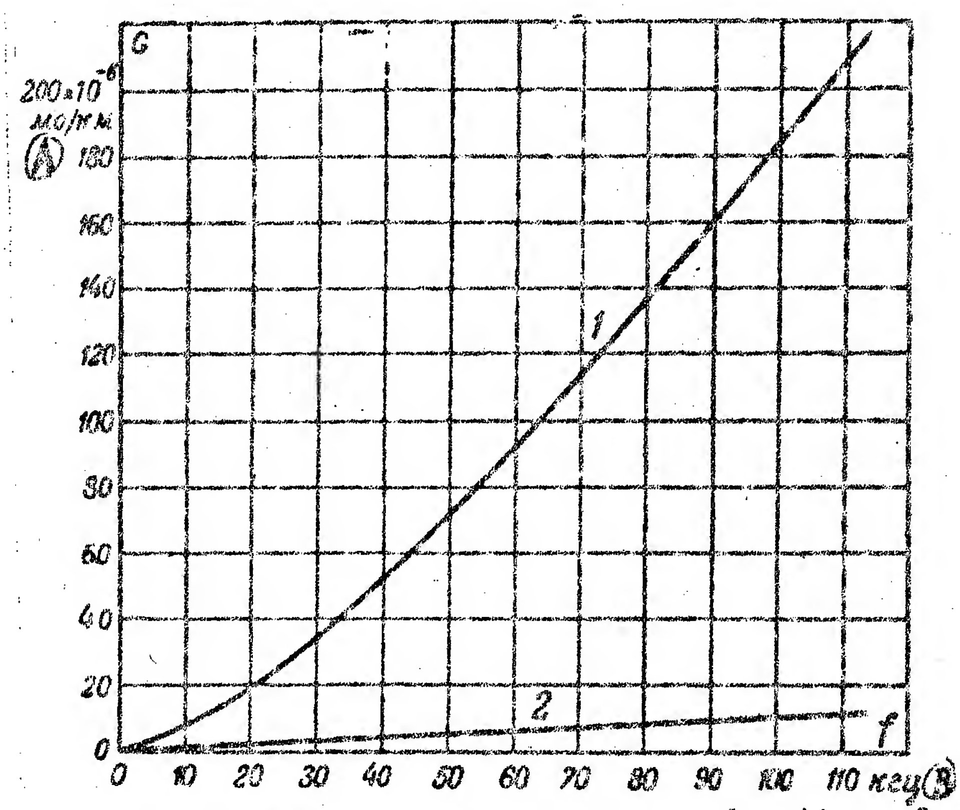


Fig. 3-30. Shunt conductance as a function of frequency.
1--paper-packthread insulation; 2--styroflex-packthread insulation.
A--mhos/km B--kcps.

The frequency-dependence of G (in the range extending to 108,000 cycles) for paper-packthread and styroflex-insulated communications cables 1.2 mm in diameter is shown in Fig. 3-30. The following cable capacitances were taken for styroflex insulation 23.5×10^{-9} fd/km. The values of tan δ are presented in Table 3-34.

It follows from Fig. 3-30 that with increasing frequency, the quantity G increases from zero to quite high values.

The shunt conductance of cables with styroflex insulation is much lower, due to the low tan δ, than that of cables with paper-packthread insulation; this is particularly noticeable at high frequencies.

Table 3-32 illustrates the variation of the quantity $G_{\sim}$ as a function of the distance a between the conductors of the cable. It has been calculated for f = 60,000 cycles and tan $\delta = 91 \times 10^{-4}$.

Table 3-32

SHUNT CONDUCTANCE AS A FUNCTION OF THE RATIO a/d.

$\frac{a}{d}$	1,5	2	3	4	5
$G_{\sim}$, μmhos/km	321	163,5	102	81,4	71,5

The decrease in $G_{\sim}$ with increasing $\frac{a}{d}$ is accounted

for chiefly by the reduction in the cable capacitance, which is a component of the parameter G.

It must be noted that the value G is a basic factor in determining the degree of high-frequency multiplexing in communications cables. High-frequency dielectrics (styroflex, polyethylene, opanol, etc.) with small dielectric loss angles have recently been widely introduced in order to expand the range of frequency utilization of communications cables.

3-8. BASIC RELATIONSHIPS FOR PRIMARY PARAMETERS OF SYMMETRICAL CIRCUITS

On the basis of the above analysis of the primary parameters of cables, we have constructed comparative characteristics for their dependence on frequency (Fig. 3-31), and the curves of variation of R, L, C, and G as functions of the diameter of the conductors and the distance between them (Fig. 3-32 and 3-33).

The orders of magnitude of the primary parameters of the most important communications cables are as follows: resistance R = (40 to 100) ohms/km, inductance L = (0.6 to 1) millihenry/km, capacitance C = (23 to 50) nanofarads/km. The shunt conductance G = (1 to 200) μmhos/km.

The following conclusion proceeds from comparison of the electrical parameters of cables and aerial communi-

cations lines. In cable circuits having relatively thin and close-packed conductors, the resistance R and the capacitance C prevail. The capacitance of the cable is 3 to 5

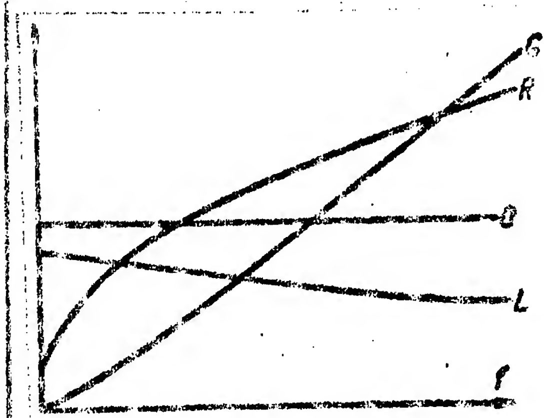


Fig. 3-31. Frequency-dependence of primary cable parameters.

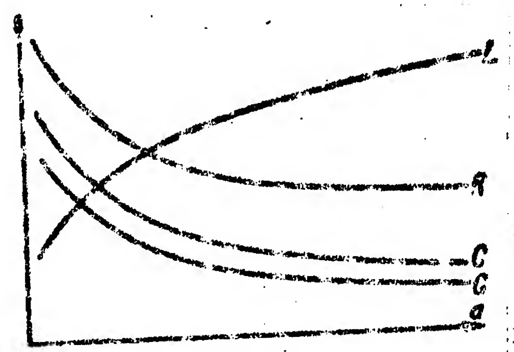


Fig. 3-32. Variation of primary cable parameters with increasing distance between conductors.

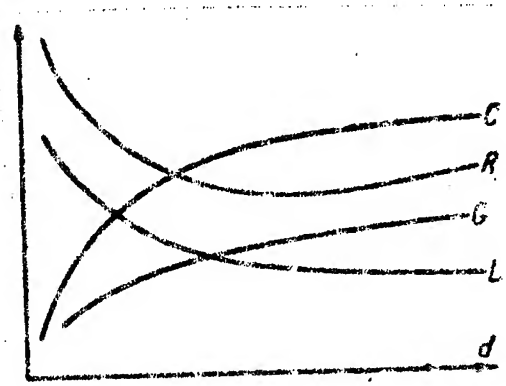


Fig. 3-33. Variation of primary cable parameters with increasing conductor diameter.

times that of the aerial line. The resistance (R) of the cable (d = 1.2 mm) is greater than the resistance of the aerial line (d = 4 mm) by a factor of approximately 15.

The inductance of the cable lines is about one-third that of the aerial lines.

3-9. CALCULATION OF DIELECTRIC CONSTANT AND DIELECTRIC LOSS FACTOR

As a rule, symmetrical communications cables have a composite insulation consisting of a dielectric and air. For the most part, paper or styroflex is used as the dielectric.

The resulting values of the dielectric constant and the dielectric loss angle of the complex insulation are determined by the electrical properties (ε and $\tan\delta$) and the volume relationships of the components forming it, with the resultant values of ε and $\tan\delta$ of the complex insulation close to the corresponding values for the parts of the insulation occupying the largest volume.

Taking into account that symmetrical cables have continuous insulation which is applied uniformly over their length, the volume relationship may be replaced by the relationship between the cross-sectional areas S. Then the resultant (equivalent) values ε_e and $\tan\delta_e$ may be computed from the formulas

д → d
в → a

$$\varepsilon_{\text{э}} = \frac{\varepsilon_{\text{д}} S_{\text{д}} + \varepsilon_{\text{в}} S_{\text{в}}}{S_{\text{д}} + S_{\text{в}}}, \qquad (3\text{-}14)$$

э → e
tg → tan

$$\operatorname{tg}\delta_{\text{э}} = \frac{\varepsilon_{\text{д}} S_{\text{д}} \operatorname{tg}\delta_{\text{д}} + \varepsilon_{\text{в}} S_{\text{в}} \operatorname{tg}\delta_{\text{в}}}{\varepsilon\,(S_{\text{д}} + S_{\text{в}})}, \qquad (3\text{-}15)$$

where the subscript-d values of ϵ and $\tan\delta$ apply to the dielectric, and the sub-a values to the air; S_d and S_a are the cross-sectional areas of the dielectric and the air.

Determination of the values of S_d and S_a presents the greatest complication, since the insulation is applied to the conductor spirally and is usually deformed in laying-up, and it is impossible to determine the fraction of the total cable section occupied by the dielectric.

For symmetrical cables with the most widely used packthread insulation, the values of S_d and S_a are determined from the expression

$$S_a = \frac{\pi D_k^2}{4} - (S_\partial + S_n). \qquad (3\text{-}16)$$

where D_c is the total diameter of the cable under the lead sheathing, and S_{cond} is the cross-sectional area of the cable conductor. In turn,

$$S_n = \frac{\pi d^2}{4} n, \qquad (3\text{-}17)$$

where d is the diameter of the cable conductor, and n is the number of conductors in the cable.

The cross-sectional area of the dielectric

$$S_\partial = \left(\pi d_1 \Delta k_1 + \frac{\pi \delta^2}{4} k_2\right) n + 2\pi D_k \Delta k_3, \qquad (3\text{-}18)$$

where Δ is the thickness of the insulation tape;

δ is the diameter of the packthread;

d_1 is the diameter of the insulated conductor;

n is the number of conductors in the cable;

k_1 and k_3 are the spiraling factors of the insulation tape;

k_2 is the spiraling factor of the packthread.

The spiraling coefficients k_1 and k_2 of the paper tape are calculated from the formula

$$k_1 = \sqrt{1 + \frac{\pi^2 (d + 2\delta + \Delta)^2}{h_p^2}}, \qquad (3\text{-}19)$$

where h_p is the lay with which the paper tape is applied.

The spiraling coefficient k_2 of the packthread is determined from the formula

$$k_2 = \sqrt{1 + \frac{\pi^2 (d + \delta)^2}{h_{th}^2}}, \qquad (3\text{-}20)$$

where h_{th} is the lay of the packthread.

It has been established by experiment that the coefficient values $k_1 = 1.17$, $k_2 = 1.32$, and $k_3 = 1.25$ may be used for the calculations.

Designing data for packthread and insulation tape are presented in Table 3-33.

Table 3-33

	Packthread-paper insulation	Packthread-styroflex insulation
Packthread δ, mm	0,4; 0,49; 0,60; 0,76; 0,85; 1,04	0,4; 0,5; 0,65; 0,8; 1,05
Tape Δ, mm	0,05; 0,08; 0,12; 0,17	0,03; 0,05; 0,20

The values of Δ and δ underlined in Table 3-33 are used in the prevailing cable designs with conductors 1.2mm in diameter.

For 32x2 cable, the volume ratios of the air, dielectric, and copper are expressed by the following figures (per unit length of cable):

$$\begin{array}{lrl}
a & V_{в} = 643 & мм^3 — 65\% \\
d & V_{д} = 75 & \text{"} \quad — 7,6\% \\
m & V_{м} = 271 & \text{"} \quad — 27,4\% \\
\hline
 & 989 & мм^3 — 100\%
\end{array}$$

In cables with packthread-paper insulation, the equivalent (resultant) value of the dielectric constant ϵ_e varies between 1.1 and 1.4 in accordance with the design of the cable. Specifically, ϵ_e for 32x2 cable intended for use for multiplexing in the range to 60,000 cycles is 1.32. In cables with paper-pulp and air-paper insulation on the conductors, the value of ϵ_e attains 1.5

to 1.7. For packthread-styroflex insulation, ϵ_e amounts to 1.2 to 1.25.

The frequency-dependence of the equivalent values tan δ_e for cables with packthread-paper and packthread-styroflex conductor insulation (d = 1.2 mm) is presented in Table 3-44.

The results of measurement of tan δ_e in a cable with packthread-paper insulation in a higher frequency spectrum are shown in Fig. 9-1.

It is clear from Table 3-34 that tan δ_e is considerably smaller for packthread-styroflex insulation than for packthread-paper insulation. This is especially pronounced in the high-frequency region; e.g., at 60 kcps it is 5, or one-eighteenth the tan δ_e of the packthread-paper insulation.

Table 3-34

Частотная зависимость $\operatorname{tg} \delta_э$ симметричных кабелей (1) с бумажно-кордельной и стирофлексно-кордельной изоляцией

(2) f, кгц	(3) Коэффициент диэлектрических потерь $\operatorname{tg} \delta_э \times 10^{-4}$		
	(4) Кабель 32×2 с бумажной изоляцией	(5) Кабель 4×4 с бумажной изоляцией	(6) Кабель 4×4 со стирофлексной изоляцией
0,8	45	35	2
5,0	49	40	2
13,5	56	45	2
20	62	50	2
30	70	60	3
40	77	70	3
50	85	80	4
60	91	90	5
70	97	100	5
80	103	110	6
90	108	115	7
100	113	120	8
108	116	125	8,5

1. FREQUENCY-DEPENDENCE OF $\tan \delta_e$ FOR SYMMETRICAL CABLES WITH PAPER-PACKTHREAD AND STYROFLEX-PACKTHREAD INSULATION
2. f, kilocycles
3. Dielectric loss factor $\tan \delta \times 10^{-4}$
4. 32x2 cable with paper insulation
5. 4x4 cable with paper insulation
6. 4x4 cable with styroflex insulation

3-10. ELECTRICAL CHARACTERISTICS OF CABLES OF SYMMETRICAL DESIGN

The MKB-32x2 is a typical representative of the group of symmetrical high-frequency cables (MKG, MPK, and

MKK)..

This cable (Fig. 3-13) is designed for cabling of main communications routes with multiplexing of circuits by 12-channel systems in the spectrum to 60,000 cycles.

In addition to the 10 high-frequency quads it contains 4 low-frequency coil-loaded quads, 3 screened radio-broadcasting circuits, and a spiral pair for monitoring the line and for auxiliary communication.

The electrical parameters of the MKB-32x2 cable are presented in Table 3-35.

Table 3-35

ELECTRICAL DATA FOR 32x2 SYMMETRICAL CABLE

Property	Test frequency,	Unit of measurement	Value	Relative length	Length conversion factor
Low-frequency quads					
1. Loop resistance	0	ohms	≤31,9	1 000	$\frac{l}{1\,000}$
2. Resistance unbalance for forward and return conductors of circuit	0	ohms	≤0,15	426	$\frac{l}{426}$
3. Insulation resistance of each conductor with reference to others and lead sheath	0(at 150v)	megohms	≥10 000	1 000	$\frac{1\,000}{l}$

4. Effective capacitance a) nominal	0,8	nfd	26,5	1 000	$\frac{l}{1\,000}$
b) permissible deviation from nominal value	0,8	nfd	±2,0	only for single-layer quads	
5. Capacitive coupling k_1	0,8	μμfd	≤120	426	$\frac{l}{426}$
k_2 and k_3	0,8	μμfd	≤300	426	$\frac{l}{426}$
6. Capacitive unbalance	0,8	μμfd	≤400	426	$\frac{l}{426}$

H i g h - f r e q u e n c y   q u a d s

1. Loop resistance	0	ohms	31,9	1 000	$\frac{l}{1\,000}$
2. Resistance unbalance of forward and return conductors of circuit.	0	ohms	≤0,10	425	$\frac{l}{425}$
3. Insulation resistance	0 (at 150v)	megohms	≥10 000	1 000	$\frac{1\,000}{l}$
4. Effective capacitance a) nominal	0,8	nfd	26,5	1 000	$\frac{l}{1\,000}$
b) permissible deviation from nominal value	0,8	nfd	+1,5 / −2,0	1 000 for all quads of second layer	
5. Capacitive coupling k_1	0,8	μμfd	≤60	426	$\frac{l}{426}$
k_2 and k_3	0,8	μμfd	≤300	426	$\frac{l}{426}$
k_9 to k_{12}	0,8	μμfd	≤20	426	$\frac{l}{426}$

Note: The coefficients k_9 to k_{12} are measured between adjacent quads.

6. Capacitive unbalance					
e_1 and e_2	0,8	μμfd	≤550	426	$\frac{l}{426}$
7. Shunt conductance	0,8	μmhos	0,85	1 000	$\frac{l}{1\,000}$
8. Magnetic coupling					
m_1	13,5	nanohenries	≤500	426	$\frac{l}{426}$
m_9 to m_{12}	13,5	нгн "	≤300	426	$\frac{l}{426}$

R a d i o - b r o a d c a s t i n g   p a i r s

1. Loop resistance	0	ohms	≤23,8	1 000	$\frac{l}{1\,000}$
2. Resistive unbalance of forward and return conductors of loop	0	ohms	0,1	426	$\frac{l}{426}$
3. Insulation resistance	0 (при at 150 в)	megohms	≥10 000	100	$\frac{1\,000}{l}$
4. Effective capacitance a) nominal	0,8	nfd	36	1 000	$\frac{l}{1\,000}$
b) permissible deviation	0,8	nfd	±3	—	—
5. Capacitive unbalance					
e_1 and e_2	0,8	μμfd	≤500	426	$\frac{l}{426}$
6. Conductance	0,8	μmhos	0,9	1 000	$\frac{l}{1\,000}$

7.Magnetic coupling between screened pairs	0,8	nanohenries	≤50	426	$\frac{l}{426}$
Over-all parameters of cable					
1.Magnetic coupling between screened pairs and layer I	0,8	nanohenries	250	426	$\frac{l}{426}$
2.Magnetic coupling between screened pairs and layer II	0,8	"	150	426	$\frac{l}{426}$
3.Crosstalk attenuation between quads					
B_1 1	60,0	nepers	≥9,0	426	—
$B_{19}-B_1$ 12	60,0	nepers	≥9,5	426	—
4.Sparkover voltage					
a)between conductors	0,05	v	1 000	—	—
b)to ground	0,05	v	1 800	—	—

The parameters of the high-frequency quads of this cable are shown in Table 3-36 as functions of frequency.

In the appendix at the end of the book, Tables A-1, A-2, A-3, A-4, A-5, and A-6 give the frequency characteristics of paired and quadded symmetrical cables with conductor diameters of 0.9, 1.2, and 1.4 mm.

Table 3-36

(1) Характеристики высокочастотных цепей 32×2 кабеля с $d = 1,2$ *мм* и бумаго-кордельной изоляцией

(2) f, кгц	(3) R, ом/км	(4) L, мгн/км	(5) C, нф/км	(6) G, мкмо/км	(7) z, ом	$-\varphi$	(8) β, мнеп/км	(9) α, рад/км
0,8	31,79	0,824		0,6	490	41°10′	43	0,049
5,0	33,25	0,824		4,08	224	23°	82,6	0,167
13,5	35,53	0,822		12,5	186	13°20′	99	0,407
20	37,7	0,821		20,8	181	9°54′	107	0,59
30	41,57	0,818		35	179	7°20′	120	0,885
40	44,91	0,815		51,5	178,5	6°	134	1,182
50	48,89	0,811	26,5	70,8	177	5°12′	141,5	1,465
60	51,6	0,807		90,8	175,5	4°32′	155	1,755
70	56,37	0,803		113,2	175	4°15′	171,5	2,03
80	59,93	0,799		137	174,4	3°58′	180	2,28
90	63,3	0,795		162	174	3°41′	197,5	2,62
100	66,7	0,792		186	173	3°30′	208,5	2,89
108	68,85	0,790		209	172,9	3°20′	218	3,12

1. CHARACTERISTICS OF HIGH-FREQUENCY CIRCUITS OF 32x2 CABLE WITH d = 1.2 mm AND PAPER-PACKTHREAD INSULATION
2. f, kilocycles
3. R, ohms/km
4. L, millihenties/km
5. C, nfd/km
6. G, μmhos/km
7. z, ohms
8. β, millihenries/km
9. α, rad/km

In calculations for screened pairs, the attenuation figures shown in these tables should be increased by the amount of the additional attenuation due to losses in the screen and the increased capacitance of the circuit. The values of this additional attenuation are presented in

Table A-7.

The electrical parameters of cables with conductors from 0.6 to 2 mm are presented in Table A-8 for a frequency of 800 cycles.

Chapter Four

CABLES WITH ARTIFICIALLY INCREASED INDUCTANCE

4-1. THE NECESSITY OF ARTIFICIAL ELEVATION OF THE INDUCTANCE OF COMMUNICATIONS CABLES

One of the pressing problems of cable technology is the extension of ranges of communication and expansion of the range of frequency utilization of the circuits without additional outlay of nonferrous metals (copper, lead) in their fabrication. This problem is being solved on the one hand by improvement of communications apparatus and on the other by reducing attenuation in the cable circuit. Below we shall consider the present methods of reducing attenuation in communications cables and give an analysis of their effectiveness in the light of current requirements.

As was indicated in the previous chapter, the electrical properties of a communications cable of any type are fully characterized by its four primary parameters; attenuation is related to these by the expression

$$\beta = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} = \beta_R + \beta_G,$$

where β_R and β_G are the attenuations of energy due to losses in the metallic parts and the dielectric of the cable, respectively.

It is perfectly clear that with R = G = 0, the transmission of energy would occur without losses, and therefore without attenuation as well. But to create such a line is impossible, since any real cable circuit possesses a resistance R and a shunt conductance G. We can only select a relationship between the circuit parameters such that its attenuation will be minimized.

Introducing the arbitrary quantity $x = \sqrt{\frac{RC}{LG}}$ into the expression given above, we obtain

$$\beta = \frac{\sqrt{RG}}{2} x + \frac{\sqrt{RG}}{2} \cdot \frac{1}{x} = \frac{\beta_0}{2} x + \frac{\beta_0}{2} \cdot \frac{1}{x},$$

where $\beta_0 = \sqrt{RG}$.

From this it is easy to show that the circuit attenuation will have a minimum ($\beta = \beta_{min}$) at x = 1, i.e., when the primary circuit parameters are linked by the relationship

$$RC = LG. \tag{4-1}$$

It is obvious that Relationship (4-1) is optimal and an effort should be made to approach it in designing communications cables. The lowest circuit attenuation attainable hereby is

$$\beta_{min} = \frac{\beta_0}{2} x + \frac{\beta_0}{2} \cdot \frac{1}{x} = \beta_0 = \sqrt{RG}. \quad (4\text{-}2)$$

Fig. 4-1 indicates the nature of the variation of the attenuation β_R in the metal and of β_G in the dielectric for various values of x. It follows from the curve that as x increases, β_R increases and β_G drops sharply. For x = 1, the losses in the metal become equal to the losses in the dielectric ($\beta_R = \beta_G$) and the cable's attenuation has its minimum value

$$\beta_{min} = \beta_R + \beta_G = \beta_0 = \sqrt{RG}.$$

In cables of existing types, x > 1, since R and C are larger in magnitude than L and G (RC ≫ LG)

It follows from the above that attenuation may be reduced either by reducing R and G, which is extremely difficult, since the values of R and G are dictated by the admissible outlay of copper (conductor diameter) and the quality of the dielectrics, or by reducing the circuit capacitance C, or by increasing its inductance L. But reduction of capacitance entails moving the conductors

farther apart, thereby increasing the dimensions; this is obviously inexpedient.

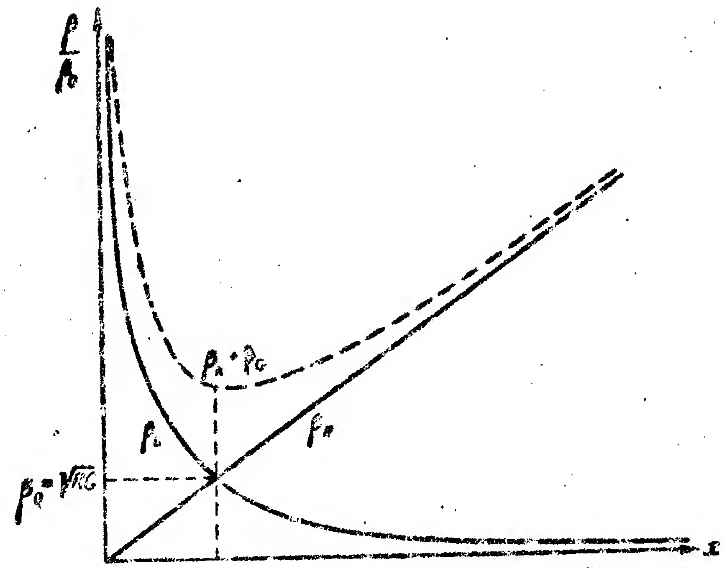
The only feasible method of reducing attenuation in cable communications lines is artificial elevation of the circuit inductance. We find from expression (4-1) that the optimal inductance value which a cable circuit must have to assure minimal attenuation is

$$L_0 = \frac{RC}{G}. \qquad (4\text{-}3)$$

The extent to which cables of the various types correspond to Relationship (4-1) is indicated on Fig. 4-2. The primary-parameter relationships of the cable in accordance with the condition $x = \sqrt{\frac{RC}{LG}}$ are laid off on the axis of abscissas, and the ratios of the actual value of the attenuation to the optimal value $\left(\frac{\beta}{\beta_0}\right)$ are plotted against the axis of ordinates. It follows from the curve that $x = 25$ in the audio-frequency range for long-distance communications cables with paper insulation. Their attenuation is accordingly $\frac{\beta}{\beta_0}$ or 13 times optimal.

Consequently it is necessary to increase the inductance of these cables artificially to reduce their parameters to Relationship (4-1).

The nonconformity of the cables to Condition (4-1) diminishes greatly with increasing frequency. Thus, for



Attenuation in metal and dielectric in various relationships of primary cable parameters.

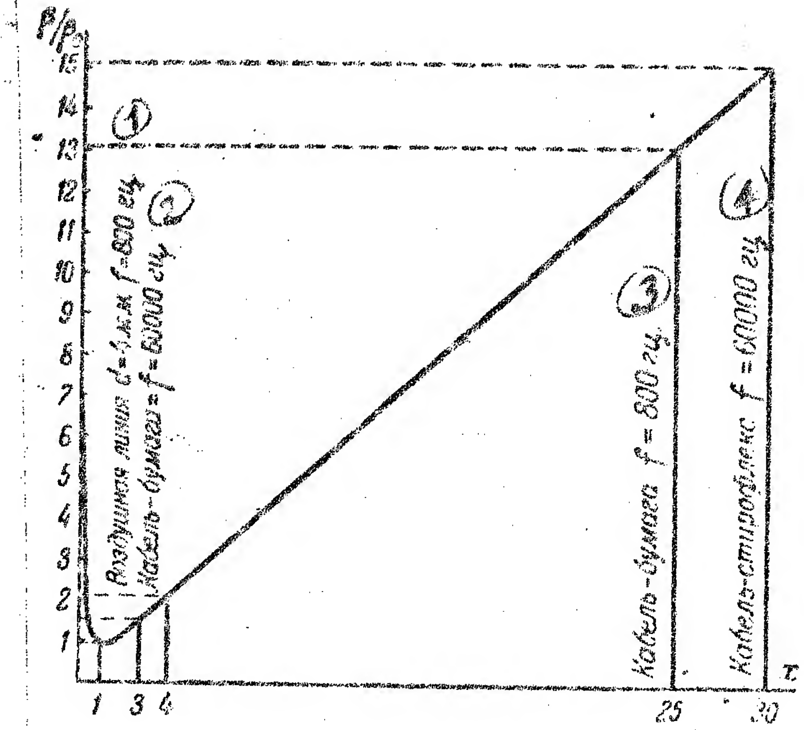


Fig. 4-2. Cable attenuation as a function of per-kilometer parameter ratios.
1--aerial line with d = 4 mm, f = 800 cps
2--cable-paper-f = 60,000 cps
3--cable-paper-f = 800 cps
4--cable-styroflex-f = 60,000 cps

example, x = 4 at 60 kcps for cables of the same type, and artificial elevation of inductance gives no significant effect.

This is accounted for by the increase in the shunt conductance G with increasing frequency, with the result that Condition (4-1) is satisfied for a certain frequency ω_x without artificially increasing the inductance. The value of ω_x may be found from the same condition (4-1):

tan

$$\frac{R}{L} = \frac{G}{C} = \frac{\omega_x C \operatorname{tg} \delta}{C} = \omega_x \operatorname{tg} \delta, \qquad (4\text{-}4)$$

from which $\omega_x = R/L \tan \delta$.

However, the frequency $f_x = \omega_x/2\pi$ lies in the range from 200 to 600 kcps for symmetrical communications cables for known types, and it is necessary to resort to artificial increases in inductance to reduce attenuation in the spectrum of practically useful frequencies. Otherwise this measure is sometimes economically impractical even at relatively low frequencies.

The thing is that as the frequency increases, Condition (4-1) is first satisfied in cables with poor dielectric (their $\tan \delta$ are large) and only much later in cables with high-quality dielectrics (their $\tan \delta$ are small). This is clear from Expression (4-4) or (4-5)

$$\frac{\omega_{x1}}{\omega_{x2}} = \frac{\mathrm{tg}\,\delta_2}{\mathrm{tg}\,\delta_1}. \qquad (4\text{-}5)$$

[tg = tan]

Thus, for example, in styroflex-insulated cable $\tan \delta = 2 \times 10^{-4}$, and the corresponding frequency ω_x is 60 times as high as for paper-insulated cable, so that in the former case it is expedient to increase inductance artificially over a rather wide frequency spectrum.

It will be seen from the curve in Fig. 4-2 that x is 30 for styroflex-insulated cable at 60 kcps, but only 4 for paper-insulated cable. Thus we may reduce the atten-

uation of a styroflex-insulated cable by a factor of nearly 15 by changing its parameter ratio from $x = 30$ to $x = 1$. It is noted in passing that at $f = 800$ cycles, the primary-parameter ratio $x \approx 1$ to 3 for 4-mm copper aerial communications lines, while $x = 25$ for paper-insulated cables. This is accounted for by the fact that the parameters R and C are significantly smaller for aerial lines than for cable lines.

An increase in the inductance of cable communications lines is expedient for yet another reason, apart from the lowered attenuation: as we know, the characteristic impedance of cables is capacitive in nature--$Z = |Z| e^{-j\omega}$. By artificially increasing the inductance, we may compensate the capacitive component and convert the complex resistance Z into a purely resistive one, $Z(\varphi = 0)$.

If such a line is loaded onto a matched impedance $Z_{pr} = Z$, the energy in the circuit will be propagated without reflection and with maximum efficiency. This assures good matching of the cable line with terminal repeater apparatus over a wide spectrum of frequencies and an accordingly maximal yield of energy to the receiver.

It is also noted that communications signals pass through lines in which Condition (4-1) is observed with

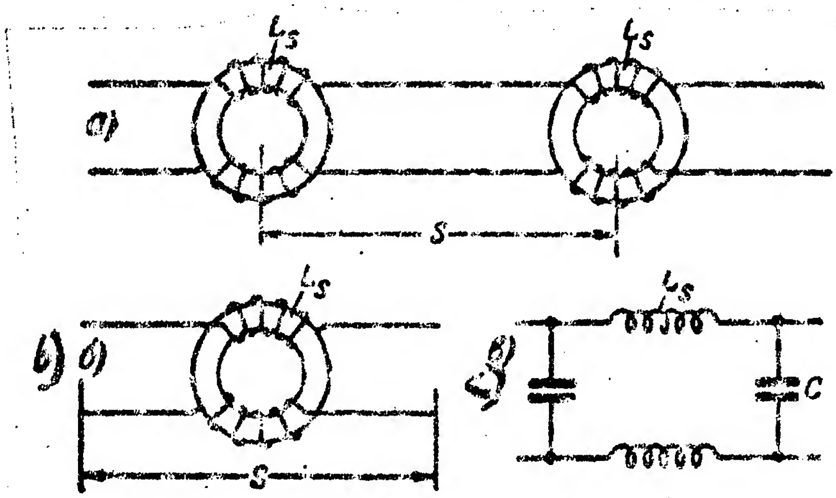


Fig. 4-3. Illustrating electrical calculation of coil-loaded cable.
a--coil-loaded circuit; b--coil-loading link
c--low-frequency filter

smaller amplitude and phase distortion, since the frequency-dependence of attenuation and velocity of propagation of energy in such lines is considerably less pronounced than in ordinary cable lines.

The secondary parameters of cable lines with artificially elevated inductance (at x = 1) are computed by the following formulas (see Chapter 2): attenuation per kilometer $\beta = \beta_O = \sqrt{RG}$, phase constant $\alpha = \omega\sqrt{LC}$, velocity of propagation $v = \omega/\alpha = 1/\sqrt{LC}$, characteristic impedance $Z = \sqrt{\frac{L}{C}}$.

Frequency does not figure in the above expressions, but the values of β, Z, and v vary somewhat with frequency due to the frequency-dependence of the primary parameters

R, L, C, and G.

Several different methods are known for artificially elevating the inductance of cable communications circuits: coil-loading, the use of ferromagnetic windings, bimetallization of the conductors, and finally the use of magneto-dielectrics.

The method of coil-loading of cable circuits has been most widely applied in cable technology.

4-2. Electrical Design of Coil-loaded Cables

Coil-loading consists in connecting coils of inductance L_S into the circuit at definite intervals called loading steps S (Fig. 4-3a).

A coil-loaded line is nonuniform in its general form, since uniform segments of the cable alternate with induction coils having quite different properties. It

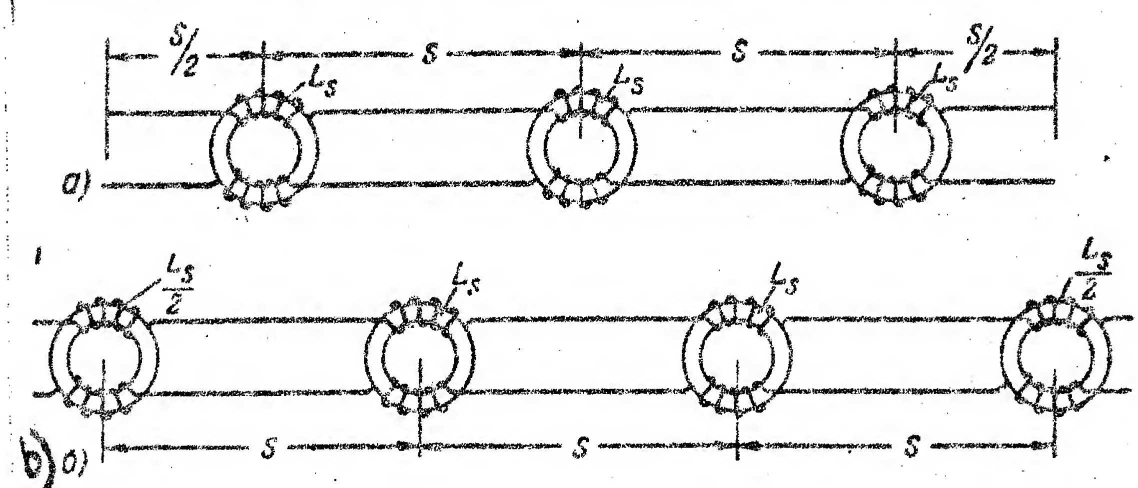


Fig. 4-4. Systems for termination of coil-loaded cables.

becomes a circuit with a periodically recurring nonuniformity when the distances between the coils are equal over the entire length of the coil-loaded line. In order for a coil-loaded circuit consisting of an integral number of identical links to present an electrically uniform system, it must be terminated at either end by either a loading half-step S/2 (Fig.4-4a) or a half-coil $L_S/2$ (Fig. 4-4b).

In existing coil-loading systems for long-distance communications cables, S is 0.285 to 1.7 km, and L_S = 1.75 to 140 millihenries. A segment of the line one S in length and including one coil L_S (Fig. 4-3b) is called a coil-loading link. Electrically, the coil-loaded line is analogous to the low-frequency filter (Fig. 4-3c). It passes a definite low-frequency spectrum (the "pass band") and holds back high frequencies (the "attenuation band"). This characteristic is a major disadvantage of coil-loaded cables. The cutoff frequency of coil-loading ω_0 may be found from the expression

$$\omega_0 = \frac{2}{\sqrt{L_{3\text{в}} C_{3\text{в}}}}, \quad (1) \qquad (4\text{-}6)$$

1--zv (link)

where L_{zv} and C_{zv} are the inductance and capacitance of the coil-loading link.

Coil-loaded circuits are calculated by the formulas which follow for the pass band from 0 to ω_0.

a) Attenuation

The attenuation of a coil-loading link

$$b = \frac{b_0 K}{\sqrt{1-\eta^2}}, \qquad (4\text{-}7)$$

where

$$K = \sqrt{\frac{2x}{x+\sqrt{x^2+1}}}, \ \underset{\text{and}}{\text{a}} \ x = \frac{\eta(1-\eta^2)}{b_0}.$$

In turn, $\eta = \frac{\omega}{\omega_0}$,

where ω and ω_0 are the theoretical and cutoff frequencies of the cable and

$$b_0 = \frac{RS\left(1-\frac{2}{3}\eta^2\right)}{2}\sqrt{\frac{C_{зв}}{L_{зв}}} + \frac{R_s}{2}\sqrt{\frac{C_{зв}}{L_{зв}}} + \frac{G_{зв}}{2}\sqrt{\frac{L_{зв}}{C_{зв}}}, \qquad (4\text{-}8)$$

[In (4-8), зв = zv = link]

where quantities with the subscript zv have reference to the coil-loading link:

$$R_{зв} = RS + R_S; \quad L_{зв} = LS + L_S;$$
$$C_{зв} = CS + C_S; \quad G_{зв} = GS + G_S;$$

[зв = zv = link]

(R, L, C, G и R_S, L_S, C_S, G_S — are accordingly parameters of the cable and coil).

The attenuation constant (per kilometer) of the coil loaded cable is determined from the expression $\beta = \frac{b}{S} = \frac{b_0 K}{\sqrt{1-\eta^2}}$,

and since in the frequency spectrum of interest to us $0{,}2<\eta<1$, parameter $K=1$, then

$$\beta = \frac{1}{2} R \frac{1-\frac{2}{3}\eta^2}{\sqrt{1-\eta^2}} \sqrt{\frac{C_{зв}}{L_{зв}}} + \frac{1}{2} \cdot \frac{R_S}{S\sqrt{1-\eta^2}} \sqrt{\frac{C_{зв}}{L_{зв}}} + \frac{1}{2} \frac{G_{зв}}{S\sqrt{1-\eta^2}} \sqrt{\frac{L_{зв}}{C_{зв}}}$$

It is more convenient to use a somewhat modified expression for calculation and analysis of attenuation in coil-loaded circuits.

Designating $\frac{1}{\eta} = \frac{\omega_0}{\omega} = \frac{2}{\omega \sqrt{L_{зв} C_{зв}}} = \nu$, we obtain after modification

$$\beta = \frac{\omega}{4} R(CS + C_S) \frac{\nu^2 - \frac{2}{3}}{\sqrt{\nu^2 - 1}} + \frac{\operatorname{tg} \varepsilon}{S\sqrt{\nu^2 - 1}} \frac{L_S}{L_S + LS} + \frac{\operatorname{tg} \delta}{S\sqrt{\nu^2 - 1}}.$$

[tg = tan]

Since the capacitance of the cable $CS \gg C_S$, the capacitance of the coil, and the former's inductance $LS < L_S$, the inductance of the coil, the expression for the attenuation of the coil-loaded circuit finally takes the form

$$\beta = \frac{\omega}{4} RCS \frac{\nu^2 - \frac{2}{3}}{\sqrt{\nu^2 - 1}} + \frac{\operatorname{tg} \varepsilon}{S\sqrt{\nu^2 - 1}} + \frac{\operatorname{tg} \delta}{S\sqrt{\nu^2 - 1}} \text{ nepers/km}, \tag{4-10}$$

[tg = tan]

where $\tan \varepsilon = R_S/\omega L_S$ and $\tan \delta = G/\omega C$ are the electrical-loss factor in the coils and the dielectric-loss factor in the cable, respectively.

As a result of the fact that $CS \gg C_S$ and $LS < L_S$, the cutoff frequency of coil-loading ω_0 may be expressed

with the same degree of approximation as

$$\omega_0 = \frac{2}{\sqrt{L_{38}C_{38}}} \approx \frac{2}{\sqrt{L_S CS}} \qquad (4\text{-}11)$$

and consequently

$$\nu = \frac{2}{\omega\sqrt{L_S CS}}.$$

It should be noted in analyzing Formula (4-10) that it consists of three parts: $\beta = \beta_R + \beta_S + \beta_G$:

of which the first, $\beta_R = \frac{\omega}{4} RCS \frac{\nu^2 - \frac{2}{3}}{\sqrt{\nu^2 - 1}}$, characterizes attenuation due to electrical losses in the conductors and other metallic parts of the cable (lead, armor, screening, etc.);

the second, $\beta_S = \frac{\tan\ \mathrm{tg}\,\varepsilon}{S\sqrt{\nu^2 - 1}}$, defines the attenuation due to electrical losses in the core of the loading coil (in eddy currents,* hysteresis, etc);

the third, $\beta_G = \frac{\tan\ \mathrm{tg}\,\delta}{S\sqrt{\nu^2 - 1}}$ denotes the attenuation due to dielectric losses in the cable insulation (dielectric losses in the loading coil may be disregarded).

b) Phase Constant

The phase displacement of the loading link is determined by the formula

$$\sin a = \frac{2\eta\sqrt{1 - \eta^2}}{K}. \qquad (4\text{-}12)$$

For the frequency spectrum in which $0.2 < \eta < 1$,

$$\sin a = 2\eta\sqrt{1-\eta^2}. \tag{4-13}$$

The phase constant (the phase shift per kilometer) is

$$a = \frac{a}{S} \; [рад/км].$$

$$\alpha = \frac{a}{S} \qquad [\text{radians/km}]. \tag{4-14}$$

c) Characteristic Impedance

The characteristic impedance Z of a coil-loaded cable depends heavily on whether the circuit is terminated in a half-step S/2 or in a half-coil $L_S/2$. If both ends of the line are terminated by half-steps, then

$$Z_{S/2} = \sqrt{\frac{L_{зв}}{C_{зв}}} \cdot \frac{1}{K\sqrt{1-\eta^2}} - j\frac{b}{2}\sqrt{\frac{L_{зв}}{C_{зв}}} \cdot \frac{1}{\eta(1-\eta^2)} \tag{4-15}$$

[зв = zv = link]

If, on the other hand, both ends terminate in half-coils,

$$Z_{L/2} = \sqrt{\frac{L_{зв}}{C_{зв}}} \cdot \frac{\sqrt{1-\eta^2}}{K} - j\frac{b}{2}\sqrt{\frac{L_{зв}}{C_{зв}}} \cdot \frac{1}{\eta}. \tag{4-16}$$

[зв = zv = link]

Figures 4-5, 4-6, and 4-7 give typical frequency curves of the attenuation β_P [P = coil-loaded] , the phase constant α_P, and the characteristic impedances $Z_{S/2}$ and

and $Z_{L/2}$ for a coil-loaded cable as compared with the analogous curves for a non-coil-loaded cable (β, α, Z).

The frequency curves of the separate attenuation components β_R, β_S, β_G (paper β_{G1}, Styroflex β_{G2}) for a coil-loaded cable are illustrated in Fig. 4-8.

It proceeds from the curves shown that coil-loading reduces attenuation by a factor of 2 to 3 over a rather wide frequency band. Only at frequencies near cutoff (ω_0) and higher does it increase rapidly and even surpass that of non-coil-loaded cables.

This demonstrates the common nature of the physical properties of coil-loaded circuits and filters.

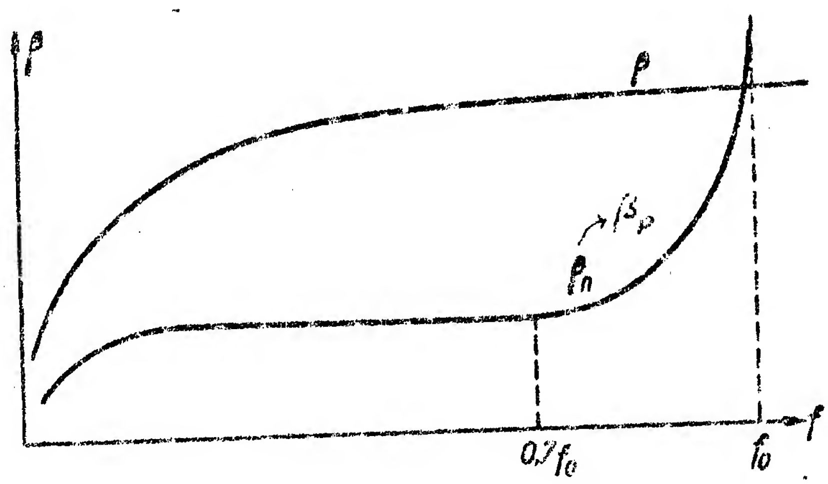


Fig. 4-5. Attenuation per kilometer of coil-loaded (β_P) and non-coil-loaded (β) cables.

Frequencies in the range from 0 to 0.7 ω_0--i.e.,

the rectilinear segment of the attenuation characteristic-- are normally used to reduce amplitude distortion in transmission over coil-loaded cables. Thus, if the highest frequency to be transmitted over the cable is ω_{max}, the coil-loading is computed for a cutoff frequency $\omega_0 = 1.43\omega_{max}$.

In evaluating the specific values of the individual components of the cable's attenuation, we may note that the attenuation β_R in the metal plays a major role. It constitutes 60-70% of the total attenuation of the cable. The attenuation β_G due to losses in the dielectric amounts to 20-30%. The remainder β_S is ascribed to attenuation in the loading coils. In cables with high-quality insula-

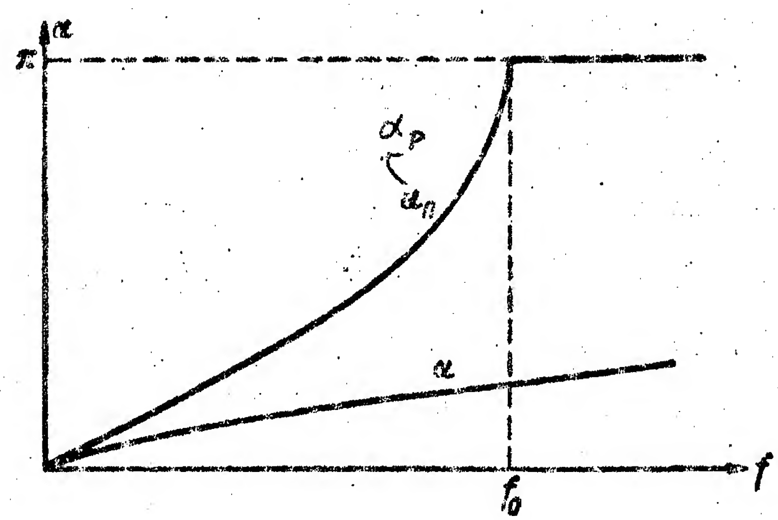


Fig. 4-6. Phase constants of coil-loaded (α_p) and non-coil-loaded (α) cables.

tion of the styroflex type, the specific value of dielectric attenuation does not exceed 1 to 3%.

It will be seen from the diagram that the value of β_C varies most sharply with increasing frequency.

The phase constant first changes comparatively slowly, almost in proportion to the increase in frequency, and then, as ω_0 is approached, rises sharply to the value π and thereafter remains constant. The phase constant of the coil-loaded cable is larger in absolute magnitude than that of the non-coil-loaded cable ($\alpha_p > \alpha$).

The characteristic impedance of a coil-loaded cable is several times larger than that on a non-coil-loaded cable. When the coil-loaded line is terminated in a half-step, the real part ("active component") of $Z_{S/2}$ goes to infinity with increasing frequency (at $\omega = \omega_0$), and $Z_{L/2}$ tends to zero in the case of termination by a half-coil.

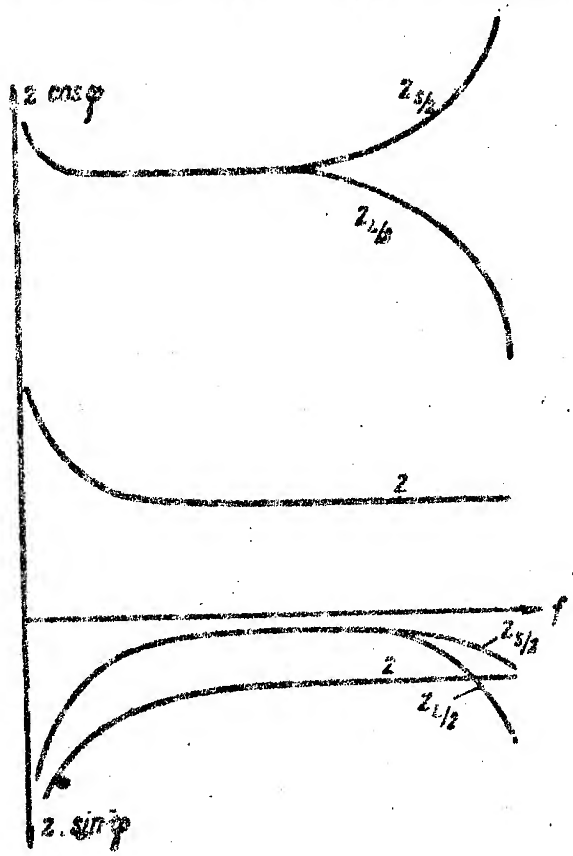


Fig. 4-7. Characteristic impedances of coil-loaded ($Z_{S/2}$, $Z_{L/2}$) and non-coil-loaded (Z) cables.

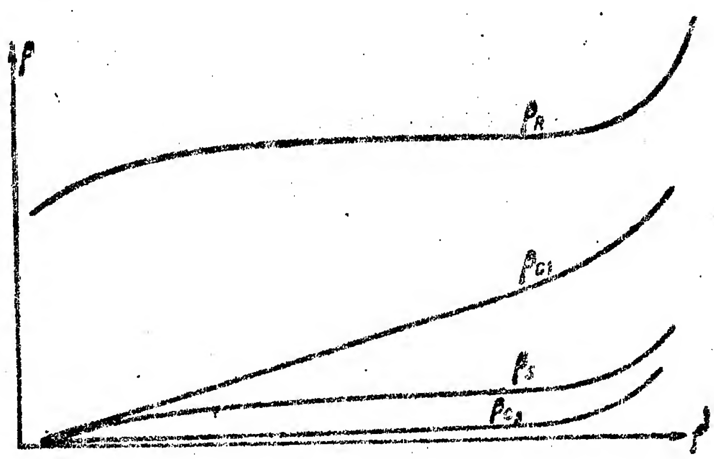


Fig. 4-8. Frequency dependence of attenuation components of coil-loaded cable.

The reactive [imaginary] component of the characteristic impedance (laid off on the negative axis of ordinates) is small compared to the active component and is no greater than 10% in absolute value.

Since the coil-loaded cable is a circuit with a periodically-recurring nonuniformity, it is extremely important that the loading links be identical to stabilize its electrical parameters. First, the parameters L and C should be as uniform as possible over the entire length of the line, and, consequently, so should the inductance of the coils and the length of the coil-loading step. In particular, so that the deviation of the value of the coil-loaded cable's characteristic impedance will not exceed ± 5%, it is necessary that a tolerance of ± 5% be maintained in the loading step and ± 1.5% in the coil inductance.

The various shipping lengths should be grouped according to the value of the capacitance C so that it will be uniform in the separate coil-loading lengths. The highest electrical uniformity of the line should be provided at the beginning of repeater sections.

4-3. Selection and Calculation of Coil-loading System.

The selection of a coil-loading system is dictated

by the problem of attaining maximum communications range in the specified frequency spectrum; here the optimum transmission conditions are to be computed for the highest frequency of the spectrum to be used. Design of the system consists in establishing the loading step S and the coil inductance L_S that provide the most favorable mode of transmission of the specified frequency spectrum coupled with minimal outlay for coil-loading of the cable. In first approximation, the electrically optimal system may be determined with the aid of:

a) the minimum-attenuation condition (4-1), from which $L_0 = \frac{RC}{G}$, and since the value of L_0 is related to the coil-loading step S and the coil inductance L_S in which we are interested by the relation (the "degree of coil loading")

$$L_0 = \frac{L_S}{S},$$

then

$$\frac{L_S}{S} = \frac{RC}{G}; \qquad (4\text{-}17)$$

b) the expression for the cutoff frequency (4-11).

It will be seen from Formula (4-17) that the optimal inductance value L_0 may be attained at various values of L_S and S. Larger values of S and L_S are economical, but cause narrowing of the passed frequency band. At low values of S and L_S the circuit passes a wide frequency band, but the cost of coil-loading the line increases rapidly

due to the small step S.

It follows from Expression (4-11) that the necessity of passing a definite frequency spectrum (from 0 to ω_0) governs the value of the product $L_S S$. The components L_S and S of the product may obviously be varied for a specified value of ω_0.

It is advantageous from an electrical standpoint to take $L_S \rightarrow \infty$ and $S \rightarrow 0$, since we then attain a large value of $L_0 = L_S/S$, but this solution is economically inefficient. The converse version ($L_S \rightarrow 0$ and $S \rightarrow \infty$) is economical, but quite unacceptable electrically, since it does not provide the required inductance ($L_0 = L_S/S \rightarrow 0$).

The practically most favorable values of S and L_S are determined by compromise between the electrical and economic optima.

The electrical optimum is found by joint solution of Expressions (4-11) and (4-17), as a result of which

$$L_S = \frac{2}{\omega_0} \sqrt{\frac{R}{G}} \tag{4-18}$$

and

$$S = \frac{2}{\omega_0 C} \sqrt{\frac{G}{R}}, \tag{4-19}$$

where $\omega_0 = 1.43\,\omega_{max}$ is the highest transmitted frequency, for which the coil-loading system is

usually calculated;

R, G, C are the parameters of the cable.

The resulting values are approximate, since they do not take into account attenuation and losses in the coils--factors which increase substantially on high-frequency multiplexing of coil-loaded circuits. Exact values of the optimal S and L_S that take into account the basic properties of the cable and the coils may be obtained, as shown by Engineer Ye. P. Arzhannikov, from the expression for attenuation in coil-loaded circuits (4-10). Investigating it first for the minimum of the coil-loading step and then for that of the coil inductance, it is easy to obtain their optimal values:

$$S_0 = \sqrt{\frac{4(\operatorname{tg}\varepsilon + \operatorname{tg}\delta)}{\omega_{\text{макс}} RC\left(\nu^2 - \frac{2}{3}\right)}}, \qquad (4\text{-}20)$$

$$L_{S0} = S_0\left[\frac{R + \omega_{\text{макс}} L \operatorname{tg}\varepsilon}{\frac{1}{3}\omega^2_{\text{макс}} RCS_0^2 + \omega_{\text{макс}}(\operatorname{tg}\varepsilon + \operatorname{tg}\delta)}\right]. \qquad (4\text{-}21)$$

[tg = tan; MAKC = max]

Fig. 4-9 shows the attenuation per kilometer β and its individual components β_R, β_S, and β_G as functions of the coil-loading step S for a constant value of L_S.

It will be seen from the diagram that the losses β_S in the loading coils decrease (due to their diminishing number) as do the losses β_G in the dielectric. However, the losses β_R in the cable conductors increase with increasing S and the coil-loading effect decreases.

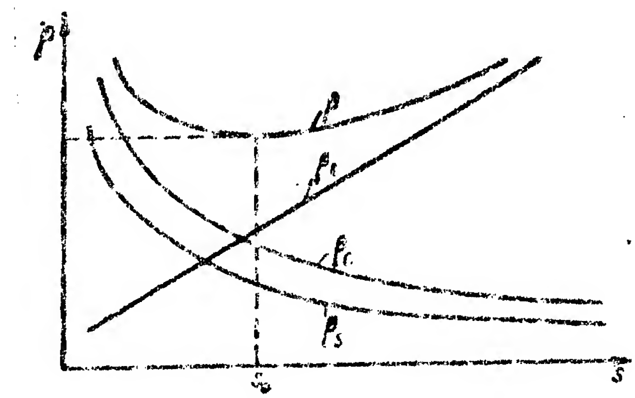


Fig. 4-9. Attenuation of coil-loaded cable as a function of loading step.

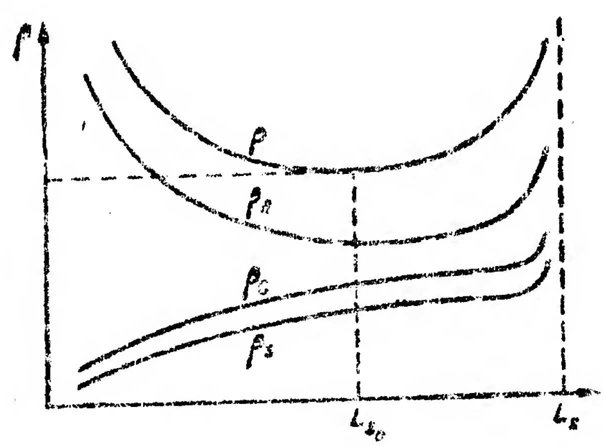


Fig. 4-10. Attenuation of coil-loaded cable as a function of loading-coil inductance with f_m and S_0 = const.

The attenuation of a coil-loaded circuit has a minimum at a definite value S_0.

Figure 4-10 shows the per-kilometer attenuation β as a function of the loading-coil inductance L_S for a given step S. It follows from the diagram that the loading in-

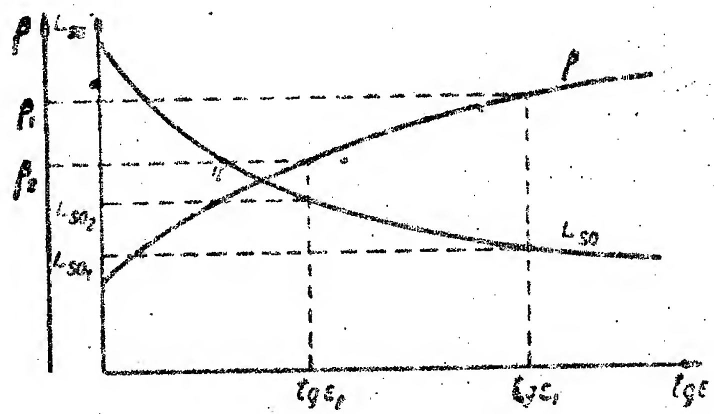


Fig. 4-11. Dependence of optimal loading-coil inductance and minimal attenuation on coil losses.

[tg = tan]

ductance also has a fully determined optimum L_{SO} corresponding to the minimal attenuation of the cable.

It follows from Formula (4-21) that the value of L_{SO} is determined not solely by the cable parameters, but also by the quality of the loading coils---their "active" resistance R_S and inductance L_S.

Figure 4-11 presents a typical curve of the optimal inductance L_{SO} of the loading coils as a function of the losses in them ($\tan \epsilon = \frac{R_S}{L_S}$). The same figure indicates

the variation of attenuation with tan ϵ. It is seen from the curve that as the losses in the coils increase, the optimal inductance value drops substantially and the attenuation of the coil-loaded circuit increases accordingly.

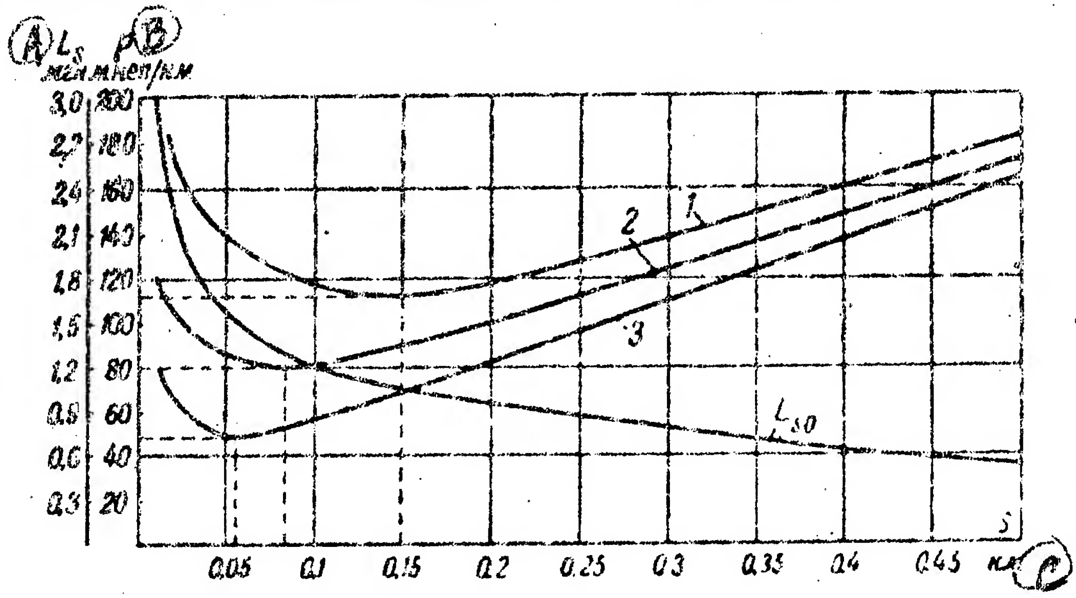


Fig. 4-12. Coil inductance and attenuation of coil-loaded cable with styroflex insulation at different values of S and tan ϵ (f = 108,000 cycles; d = 1.2 mm)
1--tan ϵ = 10.6×10^{-3}; 2--tan ϵ = 5.3×10^{-3};
3--tan ϵ = 2.65×10^{-3}.

A--L_S, millihenries. B--β, millinepers per km.
C--kilometers

This is also illustrated in Fig. 4-12, which presents the results of calculation of a cable with $d_{conductor}$ = 1.2 mm and styroflex insulation for f = 108 kilocycles and different tan ϵ for the coils.

In designing coil-loaded cables, the efforts of the

engineers are usually directed toward the attainment of minimal values of the resistive part R_S and, accordingly, of tan ϵ.

The maximum L_S for tan ϵ = 0 is expressed by the formula

$$L_{S0} = S_0 \left[\frac{R}{\frac{1}{3} \omega^2_{\text{макс}} RCS_0^2 + \omega_{\text{макс}} \operatorname{tg} \delta} \right]. \qquad (4\text{-}22)$$

[tg = tan; MAKC = max]

4-4. COIL-LOADED CABLES

As was noted in Chapter 1, coil-loaded cables belong to the category of symmetrical cables.

They are classified on the principle of frequency utilization into cables with low-frequency (ordinary) and high-frequency coil-loading. Cables with medium, light, and very light loading are low-frequency types, as are loaded radio-broadcasting cables. The high-frequency cables include frequently-loaded styroflex- and paper-insulated cables, which are suitable for multipled telephone and telegraph communication in the frequency spectrum up to 60,000 cycles.

Cable-loading systems for communications cables and

1 Современные системы пупини[illegible]

Table 4-1

Система пупинизации 2	Шаг пупинизации, км 3	Индуктивность катушки, мгн 4	Полоса передаваемых частот, гц 5	Система связи 6
			12 I. Кабели обычной [illegible]	
13 Средняя	1,7	140/56	2 400	Четырехпроводная 23
Средняя	1,7	140/56	2 400	Двухпроводная 24
Средняя	1,7	70/23	3 400	Двухпроводная
Средняя	1,7	100/70	3 400	Двухпроводная
14 Легкая	1,7	30/12	5 700	Четырехпроводная 23
15 Очень легкая . .	1,7	3,2	14 700	Четырехпроводная
16 Легкая радиовещательная . . .	1,7	12	8 000	Двухпроводная 24
			21 II. Кабели высокочастот[illegible]	
17 Частая (кабель со стирофлексной изоляцией) . .	0,285	1,75	60 000	Четырехпроводная 23
18 Частая (кабель с бумажной изоляцией)	0,425	1	60 000	Четырехпроводная 23
			22 III. Непупинизирова[illegible]	
19 Симметричный кабель	—	—	60 000	Четырехпроводная 23
20 Коаксиальный кабель	—	—	3 000 000	Четырехпроводная 23

Table 4-1 (Continued)

...изации кабелей связи

7 Диаметр жил кабеля, мм	8 Расстояние между усилителями, км	9 Количество телефонных связей	10 Дальность телефонирования, км	11 Область применения	26 Расход меди на 1 канало-километр, кг
...пупинизации					
0,9	140	1	1 400	Пригородная связь и короткие участки магистральной связи 25	15,2
1,4	140	1	1 400		18,4
1,4	100	1	1 800		—
1,2	120	1	—		
0,9	70	1+1	9 500	Дальняя связь 27	7,6
1,4	70	1+3	10 500		13,6
1,4	70	1	—	Радиовещание 28	—
...ной пупинизации					
1,2	120	1+12	6 000	Дальняя связь 27	—
1,2	55	1+12	9 000		—
...нные кабели					
1,2	35	1+12	24 500	Дальняя связь 27	3,5
2,6/9,4	14	660	28 000		1,0

[Key to Table 4-1]

1--MODERN COIL-LOADING SYSTEMS FOR COMMUNICATIONS CABLES
2--Coil-loading system
3-- Coil-loading interval, km
4--Coil inductance, millihenries
5--Transmitted-frequency band, cycles
6--Communications system
7--Diameter of cable conductors, mm
8--Distance between repeaters, km
9--Number of telephone circuits
10--Telephone range, km
11--Field of application
12--I. Ordinary loaded cables
13--Medium
14--Light
15--Very light
16--Light radio-broadcasting
17--Frequent (cables with styroflex insulation)
18--Frequent (cables with paper insulation)
19--Symmetrical cables
20--Coaxial cables
21--II. Cables with high-frequency coil-loading
22--III. Nonloaded cables
23--Four-wire
24--Two-wire
25--Suburban communications and short sections of trunk lines
26--Copper used for one channel-kilometer, kg
27--Long-distance communication
28--Radio broadcasting

---

their basic characteristics are listed in Table 4-1. It will be seen from the table that the loading interval S = 1.7 km for ordinary coil-loaded cables. The coil inductance varies and depends on the frequency range in which the circuit is used. The wider the transmission range, the smaller will be L_S.

The medium-loading system permits transmission on

one audio channel in the frequency range to 2400 cycles with a loading-coil inductance L_S = 140/56 millihenries (the numerator indicates the coil inductance for the basic circuit, and the denominator for the phantom circuit).

For telephony in the expanded frequency spectrum in use at the present time (to 3400 cycles), the loading-coil inductance L_S = 70/29 millihenries or L_S = 100/70 millihenries.

In the latter case only the spectrum to 2400 cycles passes through the phantom circuit.

The light-loading system (L_S = 30/12 millihenries) permits transmission of the frequency spectrum to 5700 cycles with multiplexing of the circuits by a single accessory channel (300 to 5700 cycles).

The very-light loading system is intended for transmission of a 14,700-cycle range; this makes it possible to form three supplementary communications channels. Here the inductance of the loading coils is only L_S = 3.2 millihenries. This loading system has not been extensively used.

The length of the repeater section on coil-loaded trunks is 2 to 4 times that of nonloaded trunks, and amounts to 140 km for the medium and 70 km for the light and very-light loading systems.

The frequency curves of attenuation in the cables are shown in Fig. 4-15 for the various coil-loading systems.

Tables 4-2, 4-3, 4-4, and 4-5 present the parameters of loaded cables.

The essential shortcoming of coil-loaded circuits it limited range of communication; here, the ultimate communication range diminishes as the degree of loading increases. This is accounted for as follows. According to the MKK (International Consultative Committee), the time for propagation of a signal from one subscriber to another must not exceed t = 250 msec if satisfactory speech quality is to be maintained.

Of these, t = 100 msec is set aside for communication between the two interurban stations. We know that the velocity of propagation of the electromagnetic energy through the wires is determined by the parameters of the circuit.

The propagation time for a signal through a 1-km section of cable line is

$$T = \frac{a}{\omega} = \frac{\omega\sqrt{LC}}{\omega} = \sqrt{LC} \quad [\text{sec/km}];$$

from which the limiting communications range is

$$l = \frac{t}{T} = \frac{100 \cdot 10^3}{\sqrt{LC}} \quad [\text{km}].$$

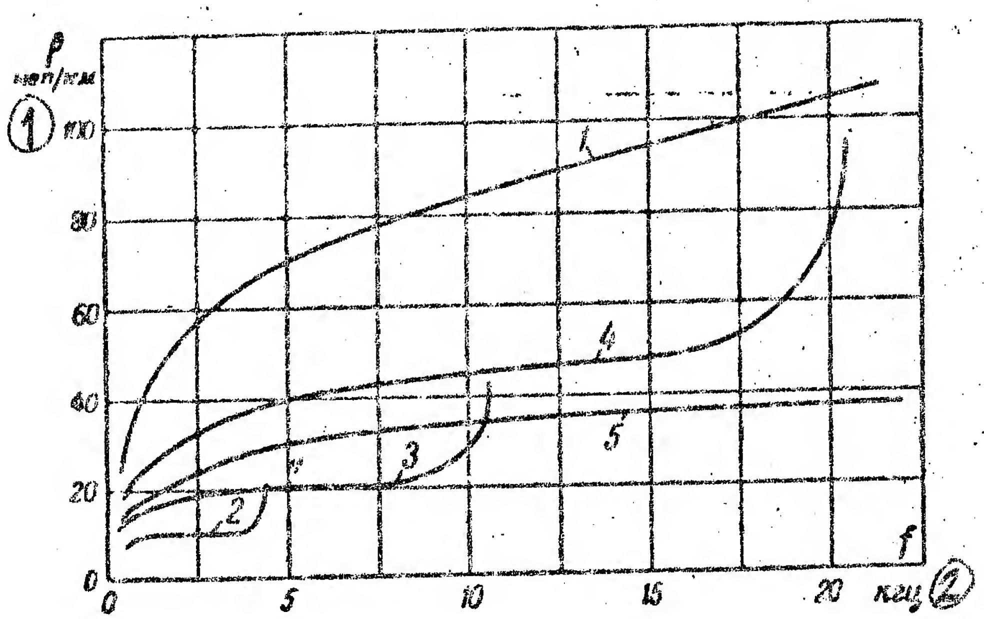


Fig. 4-13. Attenuation of cables with various coil-loading systems.

1--nonloaded cable; 2--medium coil-loading; 3--light loading; 4--very light loading; 5--frequent loading.

1--β, nepers/mm; 2--f, kilocycles

It is obvious that as the parameters L and C become larger, the signal will proceed through the line more slowly and the communication range will be shorter. It is natural that in coil-loaded cables, whose inductance is considerably higher than that of nonloaded cables, the communications range will be short. As will be seen from Table 4-1, the communications range on nonloaded cables exceeds 20,000 km; on lightly-loaded cables it drops to 3500 km, and is only 1400 km on medium-loaded cables.

Table 4-2

(1) Параметры пупинизированного кабеля

(2) $d = 0{,}9$ *мм*; скрутка жил четверкой *ДП*; $L_S = 140/56$ *мгн*; $S = 1\,700$ *м*; $t = +8°C$

(3) № по пор.	(4) f, гц	(5) Кабель длиной 1 700 м. (6) $R_{зв}$, ом	(7) $G_{зв}$, мкмо	(8) Катушка (9) R_S, ом	(10) β, мнеп/км	(11) При окончании полушагом (12) $z\cos\varphi$, ом	(13) $-z\sin\varphi$, ом
(14) Основная цепь: $L_S = 140$ *мгн*; $C = 0{,}0335$ *мкф*							
1	0	87,72	0	7,8	—	—	—
2	800	87,72	0,973	8,9	18,5	1 600	120
3	1 600	87,85	2,45	10,4	19,1	1 750	60
(15) Фантомная цепь: $L_S = 56$ *мгн*; $C = 0{,}054$ *мкф*							
4	0	43,86	0	3,87	—	—	—
5	800	43,90	1,559	4,31	18,2	800	70
6	1 600	44,05	3,15	4,88	19,0	840	30

1) PARAMETERS OF COIL-LOADED CABLE
2) d = 0.9 mm; quadded conductors (DP); L_S = 140/56 millihenries; S = 1700 m; t = +8°C.
3) No.
4) f, cycles
5) 1700-m cable
6) R_{link}, ohms
7) G_{link}, micromhos
8) Coil
9) R_S, ohms
10) β, millinepers/km
11) With terminal half-step
12) z cos φ, ohms
13) -z sin φ, ohms
14) Basic circuit: L_S = 140 millihenries; C = 0.0335 μfd
15) Phantom circuit: L_S = 56 millihenries; C = 0.054 μfd

Table 4-3

(1) Параметры пупинизированного кабеля

$d = 1{,}4$ *мм*; скрутка жил четверкой *ДП*; $L_S = 140/56$ *мгн.*; $S = 1\,700$ *м*; $t = +8°C$ (2)

(3) № по пор.	(4) f, *кгц*	(5) Кабель длиной 1 700 *м*		(8) Катушка	(10) β, *мнеп/км*	(11) При окончании полушагом	
		(6) $R_{зв}$, *ом*	(7) $G_{зв}$, *мкмо*	(9) R_S, *ом*		(12) $z\cos\varphi$, *ом*	(13) $-z\sin\varphi$, *ом*
(14) Основная цепь: $L_S = 140$ *мгн*; $C = 0{,}0355$ *мкф*							
1	0	36,21	0	7,80	—	—	—
2	0,8	36,28	1,031	8,80	9,2	1 550	50
3	1,5	36,39	2,048	10,07	10,1	1 580	30
4	2,5	36,57	3,602	12,50	12,8	2 240	30
(15) Фантомная цепь: $L_S = 56$ *мгн*; $C = 0{,}0575$ *мкф*							
5	0	18,11	0	3,90	—	—	—
6	0,8	18,20	1,671	4,33	9,0	775	35
7	1,5	18,29	3,317	4,80	9,7	815	20
8	2,5	18,49	5,835	5,60	11,2	980	20

1) PARAMETERS OF COIL-LOADED CABLE
2) d= 1.4 mm; quadded conductors (DP); L_S = 140/56 millihenries; S = 1700 m; t = +8°C
3) No.
4) f, kilocycles
5) 1700-m cable
6) R_{link}, ohms
7) G_{link}, micromhos
8) Coil
9) R_S, ohms
10) β, millinepers/km
11) with terminal half-step
12) $z\cos\varphi$, ohms
13) $-z\sin\varphi$, ohms
14) Basic circuit: L_S = 140 millihenries; C = 0.0355 μfd
15) Phantom circuit: L_S = 56 millihenries; C = 0.0575 μfd

Table 4-4

(1) Параметры пупинизированного кабеля

d (2) $= 0,9$ мм; скрутка жил четверкой ДП; $L_S = 30/12$ мгн; $S = 1\,700$ м

$t = +8°C$

(3) № по пор.	(4) f, кгц	(5) Кабель длиной 1 700 м: (6) $R_{уз}$, ом	(7) $G_{уз}$, мкмо	(8) Катушка: (9) R_S, ом	(10) β, мнеп/км	(11) При окончании полушагом: (12) $z \cos \varphi$, ом	(13) $-z \sin \varphi$, ом
(14) Основная цепь: $L_S = 30$ мгн; $C = 0,0335$ мкф							
1	0	87,72	0	5,24	—	—	—
2	0,8	87,72	0,973	5,36	35,0	770	220
3	3,0	87,98	4,079	5,92	37,5	800	70
4	5,7	88,51	8,566	7,35	41,0	1 100	50
(15) Фантомная цепь: $L_S = 12$ мгн; $C = 0,054$ мкф							
5	0	43,86	0	2,62	—	—	—
6	0,8	43,90	1,569	2,70	35,0	400	135
7	3,0	44,21	6,576	2,93	37,0	395	40
8	5,7	44,73	13,808	3,65	39,0	470	25

1) PARAMETERS OF COIL-LOADED CABLE
2) d = 0.9 mm; quadded conductors (DP); L_S = 30/12 microhenries; S = 1700 m; t = +8°C
3) No.
4) f, kilocycles
5) 1700-m cable
6) R_{link}, ohms
7) G_{link}, micromhos
8) Coil
9) R_S, ohms
10) β, millinepers/km
11) With terminal half-step
12) z cos φ, ohms
13) -z sin φ, ohms
14) Basic circuit: L_S = 30 millihenries; C = 0.0335 μfd
15) Phantom circuit: L_S = 12 millihenries; C = 0.054 μfd

Table 4-5

(1) Параметры пупинизированного кабеля

(2) $d = 1{,}4$ *мм*; скрутка жил парная *П*; $L_S = 3{,}2$ *мгн*; $S = 1\,700$ *м*; $C = 0{,}0355$ *мкф*; $t = +8°C$

(3) № по пор.	(4) f, кгц	(5) Кабель длиной 1 700 м: (6) $R_{зв}$, ом	(7) $G_{зв}$, мкмо	(8) Катушка: (9) R_S, ом	(10) β, мкеп/км	(11) При окончании полушагом: (12) z cos φ, ом	(13) — z sin φ, ом
1	0	36,21	0	1,02	—	—	—
2	1,8	36,35	2,46	1,05	39,0	285	95
3	3,0	36,61	4,32	1,07	41,0	277	60
4	6,0	37,44	10,03	1,16	42,5	280	30
5	12,0	39,83	22,75	1,39	45,8	312	20

1) PARAMETERS OF COIL-LOADED CABLE
2) d = 1.4 mm; paired conductors (P); L_S = 3.2 millihenries; S = 1700 meters; C = 0.0355 μfd; t = +8°C
3) No.
4) f, kilocycles
5) 1700-m cable
6) R_{link}, ohms
7) G_{link}, micromhos
8) Coil
9) R_S, ohms
10) β, millinepers/km
11) With terminal half-step
12) z cos φ, ohms
13) -z sin φ, ohms

<u>Note</u> to Tables 4-2, 4-3, 4-4, and 4-5: R_{link} and R_S are, respectively, the ohmic resistances of a cable circuit of length S (S being the length of the loading link or step) and the coil; G_{link} is the shunt conductance of a cable circuit of length S.

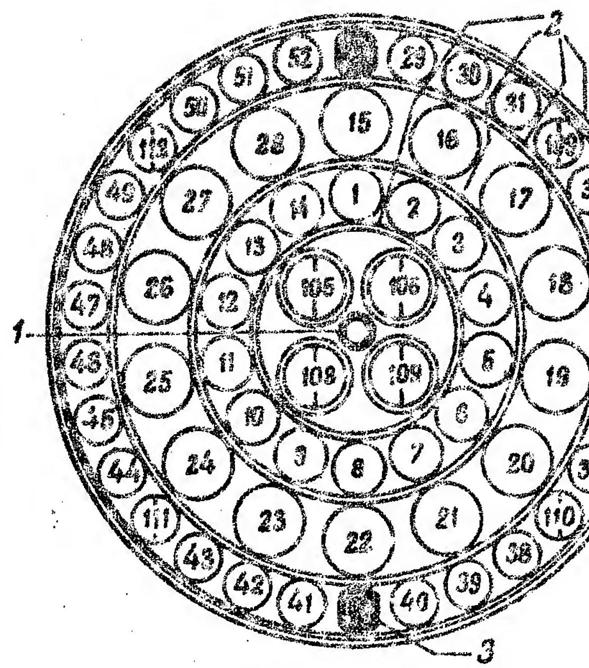
The economic expediency of using a given loading system is determined to a significant degreee by the copper outlay for one channel-kilometer (see last column of Table 4-1). The utilization factor for the quads in the low-frequency loading systems is comparatively small. Three circuits may be obtained on two quads with medium loading, and six with light loading (in the four-wire system, taking phantom circuits into account).

Thus to extablish a large number of circuits on truck routes it becomes necessary to use cables with capacities as high as 218 pairs (218 x 2).

Figure 4-14 shows a typical construction of a 112x2 cable with aluminum conductors for the ordinary loading system. Its individual circuits are loaded in accordance with their functions and the required communications range. Four screened pairs (105-108) for radio broadcasting are located in the center of the cable ($d_{cond.}$ = 1.8 mm).

Fourteen quads (d_{cond} = 1.15 mm) are placed in the first concentric layer; of these, 1, 2, 8, and 14 are medium-loaded and the remainder light-loaded; the quads 3-7 and 9-13 are used for transmission in opposite directions.

In the second layer, all pairs in the quads 15-28 (d_{cond} = 1.8 mm) are medium-loaded and used for audio-



Symbols:
pairs
quads
screened pairs

Fig. 4-14. Construction of 112x2 cable.

frequency communications.

In the third layer, the 24 quads 29-112 are very lightly loaded.

The quads 29-40 and 41-52 are intended for transmission in opposite directions. The conductors of the circuits with medium and light loading are spiraled in the DP scheme.

With full multiplexing, we may obtain the following

from such a cable: 4 radio broadcasting channels, 102 light-loaded circuits, 8 very-lightly loaded circuits, and 54 medium-loaded circuits.

The following equivalent conductor diameters are used in similar cable constructions with copper conductors:

Copper	Aluminum
1.4 mm	1.8 mm
1.2 mm	1.55 mm
0.9 mm	1.15 mm

4-5. Cables with High-frequency Loading.

Modern communications technique is developing in the direction of expanded spectra of efficiently-transmitted frequencies and increased communications ranges.

As applied to coil-loaded cables, these requirements are found to be quite contradictory, since enlargement of the transmitted-frequency band involves a lower degree of loading of the cable, with the result that the specified increase in range cannot be achieved. With high-frequency multiplexing of ordinary paper-insulated cable, the coil inductance is so small that the attenuation-reduction effect thus obtained does not justify the expense of loading the cable mains in many cases.

Therefore the tendency toward broad-band multiplex-

ing of communications lines has resulted in practical abandonment of coil-loading for paper-insulated cables and even of artificial means of increasing the inductance of cable circuits of this type. Only with the creation of high-frequency dielectrics have new prospects been uncovered for increasing communications ranges and expanding the transmitted-frequency band in coil-loaded cables . Thus, for example, styroflex-insulated, frequency-loaded cables permit of high-frequency multiplexing of the circuits in the range extending to 60,000 cycles, coupled with a two-thirds reduction of attenuation below that of nonloaded paper-insulated cables. While the repeater section is 35 km long on ordinary cable trunks, the use of the styroflex cables increases it to 120 km (d_{cond} = 1.2 mm).

It should be noted, however, that the use of the new dielectrics in symmetrical long-distance communications cables is efficient only when the circuits are coil-loaded, i.e., when their primary parameters are reduced to Relationship (4-1). Replacement of paper insulation by styroflex insulation in symmetrical cables does not in itself give a marked reduction in attenuation, and is hardly economically expedient for the frequency spectrum transmitted over it (to 60 or 108kc).

The merits of frequency-leaded cable with styroflex insulation are accounted for as follows. For pack-

thread insulation $\tan\delta = 120 \times 10^{-4}$ at a frequency of 60 kc while for styroflex-packthread insulation $\tan\delta$ is only 2×10^{-4}, or one-sixtieth as large. As a result, the 60-kc attenuation of the styroflex cable may be reduced by a factor of 7 to 8:

$$\frac{\beta_{0\,бум}}{\beta_{0\,стир}} = \frac{\sqrt{RG_{бум}}}{\sqrt{RG_{стир}}} = \sqrt{\frac{R\omega C \operatorname{tg}\delta_{бум}}{R\omega C \operatorname{tg}\delta_{стир}}} = \sqrt{\frac{\operatorname{tg}\delta_{бум}}{\operatorname{tg}\delta_{стир}}} = \sqrt{\frac{120\cdot 10^{-4}}{2\cdot 10^{-4}}} = 7{,}7. \qquad (4\text{-}23)$$

[$бум$ = paper; $стир$ = styroflex; tg = tan]

However, the value of $x = \sqrt{\frac{RC}{LG}} = 30$ (Fig. 4-2) in styroflex-insulated cables and the actual attenuation is 15 times optimal ($\beta_0 = \sqrt{RG}$). The artificial introduction of inductance changes the relation between the cable parameters, lowers the value of x and makes the attenuation of the cable approach the minimal value $\beta_0 = \sqrt{RG}$.

In the loading system adopted for styroflex cable, the coil inductance $L_3 = 1.75$ millihenries, the loading interval $S = 285$ m, the value of x is brought up to 9.5 and the attenuation is reduced by two thirds (β/β_0 is reduced from 15 to 5). By increasing the degree to which the styroflex cable is loaded, we may approach even closer

to a 7-8-fold reduction of attenuation (without considering the losses in the loading coils), but this is not justified economically.

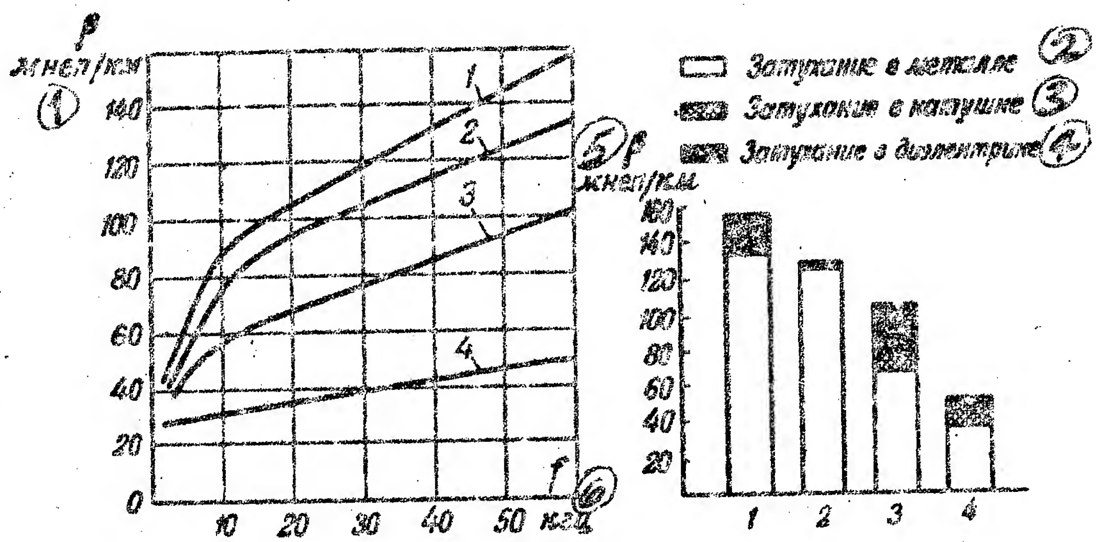


Fig. 4-15. Attenuation of cables with high-frequency loading.

1--nonloaded paper-insulated cable; 2--nonloaded styroflex-insulated cable; 3--loaded paper-insulated cable; 4--loaded styroflex-insulated cable.

1) β, millinepers/km
2) Attenuation in metal
3) Attenuation in coil
4) Attenuation in dielectric
5) β, millinepers/km
6) f, kilocycles

Figure 4-15 and Table 4-6 show frequency characteristics of the attenuation components for paper- and styroflex-insulated cables at 60,000 cycles, with coil-loading and without it. The conductor diameter is 1.2 mm. The

attenuation values for nonloaded cables with paper (1) and styroflex (2) insulation confirm that the latter gives no appreciable

Table 4-6

Составляющие затухания в мнеп/км в различных типах кабелей при частоте 60 000 гц. (1)

Тип кабеля (2)	β_R	β_S	β_G	(3) β общее
(4) Непупинизированный кабель с бумажной изоляцией	134,5		18,5	153,0
Пупинизированный с бумажной изоляцией (5)	64,6	6,8	30,6	102,0
Непупинизированный со стирофлексной изоляцией (6)	130,0	—	0,2	130,2
Пупинизированный со стирофлексной изоляцией (7)	41,2	9,0	0,8	51,0

1) ATTENUATION COMPONENTS (in millihenries/km) FOR CABLES OF VARIOUS TYPES AT A FREQUENCY OF 60,000 CYCLES
2) Type of cable
3) β (total)
4) Nonloaded paper-insulated cable
5) Loaded paper-insulated cable
6) Nonloaded styroflex-insulated cable
7) Loaded styroflex-insulated cable

effect without the use of artificial inductance. Comparison of Curves 1 and 3 testifies that for paper insulation, coil-loading reduces attenuation by only 40 to 50%. It follows from Curves 1 and 4 that at a frequency of 60 kc,

coil-loading combined with the use of styroflex insulation reduces the attenuation by a factor of three (from 153 millihenries per km for nonloaded cable with paper insulation to 51 millihenries/km for frequently-loaded styroflex cable).

Frequently-loaded paper-insulated cable ($S = 425$ m, $L_S = 1$ millihenry) have not been accorded extensive application.

Type TZSB-4x4-1.2 styroflex-insulated cable (Fig. 4-16) consists of four quads, each of which is spiraled. The conductors are insulated with styroflex packthread and tape. Paper strip insulation is applied over this. For convenience in assembly, the conductors are distinguished by the color of the packthread, and the fours are wound with colored triacetate thread. The protective lead shield is armored.

The loading coils are assembled in small boxes.

The design data of Type TZSG-4x4-1.2 cable are given in Table 4-7. Cables with 3x4 and 7x4 capacities are also manufactured.

The lays of the quads and the electrical data of styroflex-insulated cable are listed in Tables 4-8 and 4-9. The characteristic impedances of closely loaded cables with styroflex and paper insulation are shown in

Fig. 4-7.

The limiting communications range on close-loaded styroflex-insulated cable is considerably greater than that on cable with low-frequency loading, and reaches 6000 km with t = 100 msec (Table 4-1).

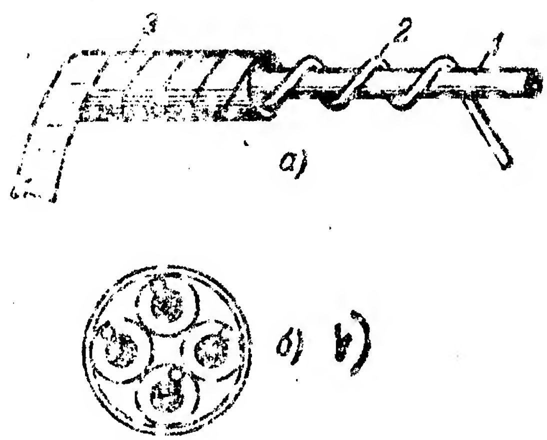


Fig. 4-16. Design of TZS-4x4-1.2 styroflex-insulated cable.

a) Cable conductor with Styroflex insulation. 1--conductor (d = 1.2 mm); 2--styroflex packthread; 3--styroflex tape; b) Spiral-four form of styroflex-insulated cable.

Figure 4-18 shows the attenuation of cables with paper and styroflex insulation, without the use of coil-loading, in the frequency spectrum to 300,000 cycles. It will be seen from the diagram that in the high-frequency

Table 4-7

(10) Конструктивные данные кабеля ТЗСГ-4 × 4.

№ по пор. (11)	Наименование (12)	Конструкция или радиальная толщина, мм (13)	Наружный диаметр, мм (14)	Вес, кг/км (15)
1	Медная жила	1×1,2	1,2	168
2	Спиральная обмотка стирофлексным корделем δ=0,8 мм с шагом 6—8 мм	0,8	2,8	11,5
3	Обмотка стирофлексной лентой размером (10—12)×0,05 мм с положительным перекрытием 25—35%	0,05	2,9	10,5
4	Скрутка четырех жил со спиральной обмоткой шелком. Шаг обмотки 19—22 мм	—	7,1	0,8
5	Общая скрутка четырех четверок с шагом 350±10 мм	—	15,5	—
6	Поясная изоляция из четырех лент кабельной бумаги толщиной 0,18 мм	0,5	16,6	26,5
7	Свинцовая оболочка с присадкой сурьмы в количестве 0,4—0,6%	1,4	19,4	945,2 4,8
	Итого . . . . (8)			1 168 кг (9)

1) Copper filament
2) Spiral winding of styroflex packthread with δ = 0.8 mm and a lay of 6-8 mm
3) Winding of styroflex tape, (10 to 12) x 0.05 mm with a plus overlap of 25-35%
4) Four spiraled conductors with spiral winding of silk. Step of winding 19-22 mm
5) Final spiraling of four quads, lay = 350 ± 10 mm
6) Strip insulation of four cable-paper tapes 0.18 mm thick
7) Lead sheathing with admixture of 0.4 to 0.6% of antimony
8) Total
9) kg
10) DESIGN DATA FOR TZSG-4x4 CABLE
11) No.
12) Designation
13) Configuration or radial thickness, mm
14) Outside diameter, mm
15) Weight, kg/km

Table 4-8

(1) **Шаги скрутки и шаги наложения цветной нити в кабеле ТЗСБ — 4×4**

(3) Элемент	(4) Шаги скрутки, мм	(5) Шаг наложения цветной нити, мм
(2) Четверка № 1	160 ± 5	(6) Красная 17,0 . . . .
№ 2	200 ± 6	(7) Синяя 21,5 . . . .
№ 3	175 ± 5	(8) Зеленая 19,0 . . . .
№ 4	125 ± 5	(9) Желтая 15,0 . . . .

1) SPIRALING LAYS AND LAYS OF APPLICATION OF COLORED THREAD IN TZSB 4×4 CABLE
2) Quad
3) Element
4) Spiraling Lay, mm
5) Lay of application of colored thread, mm
6) Red 17.0
7) Blue 21.5
8) Green 19.0
9) Yellow 15.0

Note: The quads are spiraled left-hand. The final twist of the cable is right-handed with a lay of 350 ± 10mm

[Key to Table 4-9 –next page]

1) Loop resistance
2) Difference between resistances of conductors in pair
3) Insulation resistance
4) Effective capacitance
   a) rated
   b) maximum deviation
5) Capacitive coupling
   k_1 maximum
   k_1 average
   k_2-k_3 maximum
   k_9-k_{12} maximum
   e_1-e_2 maximum
6) Modulus of magnetic coupling within and between quads
7) Test voltage
8) Index
9) Frequency, kc
10) Unit of measurement

[Key continued on Page 149]

Table 4-9

Электрические показатели кабеля со стирофлексной изоляцией марки ТЗСБ — 4 × 4 — 1,2 при $t = 20°C$

№ по пор.	Наименование показателей	Частота, кгц	Единица измерения	Норма	Длина, к которой относится норма, м	Коэффициент пересчета: для длины l, м	Коэффициент пересчета: для температуры
1	Сопротивление петли	0	ом/км	31,9	1000	$l/1000$	1,004
2	Разница сопротивления жил в паре	0	ом	0,12	285	$l/285$	
3	Сопротивление изоляции	0 х120 в	мгом	10 000	1 000	$1000/l$	
4	Рабочая емкость						
	а) номинальная	0,8	нф	23,5	1 000	$l/1000$	
	б) максимальное отклонение	0,8	нф	± 1,2	1 000	$l/1000$	
5	Емкостная связь						
	k_1 максимальная	0,8	пф	50	285	$l/285$	
	k_1 средняя	0,8	пф	25	»	$\sqrt{l/285}$	
	$k_2 - k_3$ максимальная	0,8	пф	150	»	$l/285$	
	$k_9 - k_{12}$ максимальная	0,8	пф	20	»	$l/285$	
	$e_1 - e_2$ максимальная	0,8	пф	300	»	$l/285$	
6	Модуль магнитной связи внутри четверок и между четверками	13,5	нгн	250	285	$l/285$	
7	Испытательное напряжение	0,05	эффект. вольт в течение 2-х мин.	1 800			

Примечания:

1. Модуль магнитной связи вычисляется $M = \sqrt{|m|^2 + |12r|^2}$, где r в мом, m в нгн.
2. В зависимости от величины средней рабочей емкости кабеля делятся на 8 групп

(27) Номер группы	(28) Средняя рабочая емкость всех пар в одной строительной длине (*нф/км*)
	22,70–22,90
II	22,91–23,10
III	23,11–23,30
IV	23,31–23,50
V	23,51–23,70
VI	23,71–23,90
VII	23,91–24,10
VIII	24,11–24,30

[Key to Table 4-9 (page 148), continued]

11) ELECTRICAL INDICES OF TYPE TZSB-4x4-1.2 STYROFLEX-INSULATED CABLE AT t = 20°C.
12) No.
13) Norm
14) Length to which norm refers, m
15) Conversion factor
16) For length l, meters
17) For temperature
18) x 120 v (?)
19) ohms/km
20) ohms
21) megohms
22) nanofarads
23) μμfd
24) nanohenries
25) effective volts for 2 minutes
26) Notes:
    1. The magnetic coupling modulus is computed as $M = \sqrt{|m|^2 + |12r|^2}$, with r in mohms and m in nanohenries.
    2. The cables are classed into eight groups in accordance with the value of the average effective capacitance:

27) Number of group
28) Average effective capacitance of all pairs in one shipping length (nanofarads per kilometer)

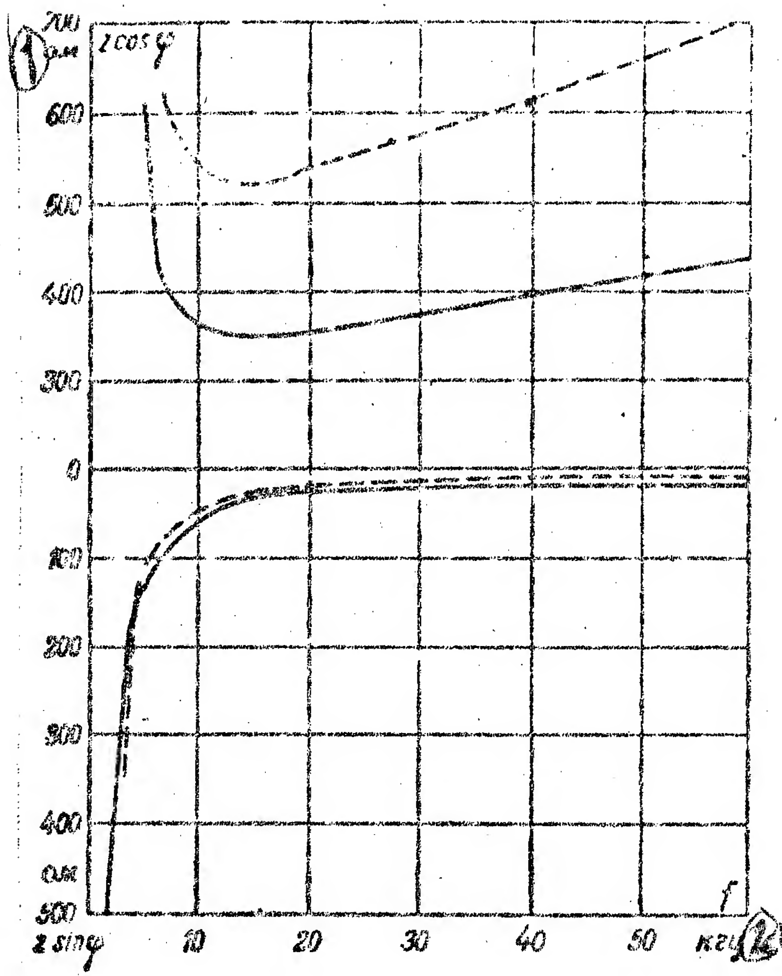


Fig. 4-17. Characteristic impedance of cables with high-frequency loading.

_ _ _ _ _ _ styroflex insulation;

____________ paper insulation.

1) ohms
2) f, kilocycles

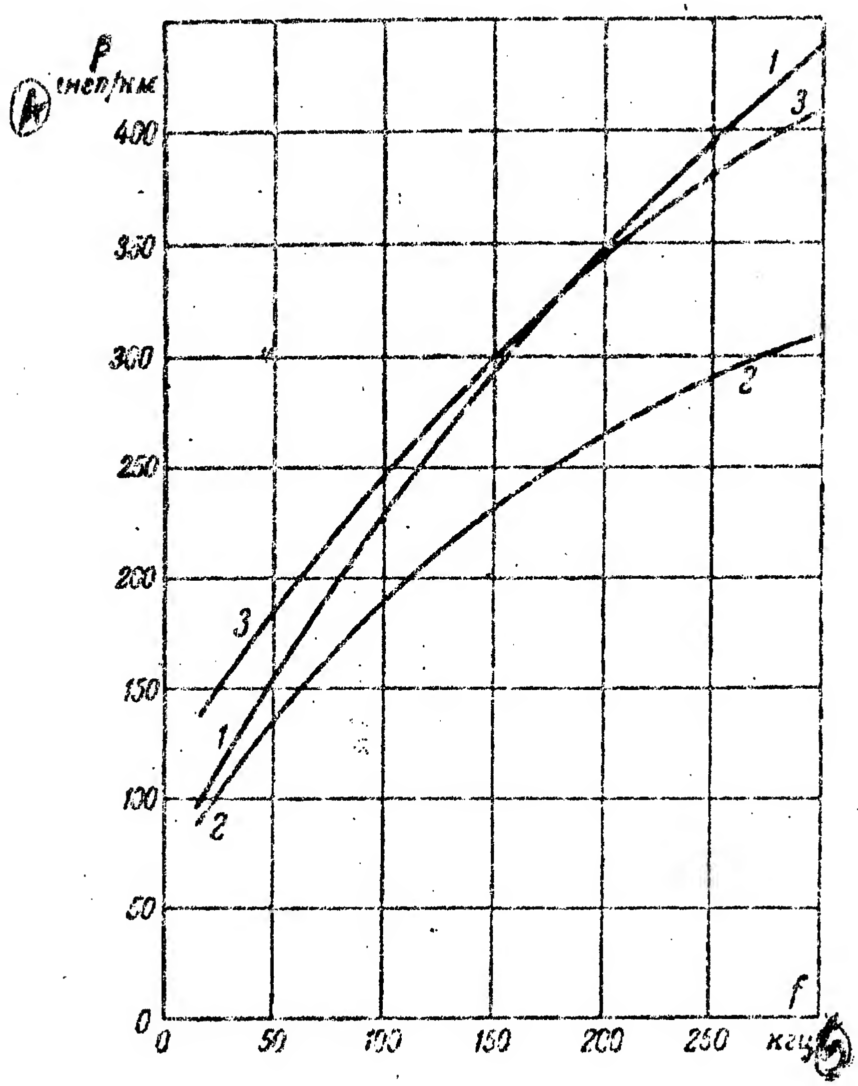


Fig. 4-18. Attenuation of unloaded cables with paper and styroflex insulation.

1) d = 1.2 mm, paper insulation; 2) d = 1.2 mm, styroflex insulation; 3) d = 0.9 mm, styroflex insulation.

4-- , [milli] nepers per km
5--f, kilocycles

region (250 kc and higher) the use of styroflex becomes efficient even without coil-loading.

At a frequency of 180 kc, the attenuation of styroflex-insulated cable with conductors 0.9 mm in diameter is

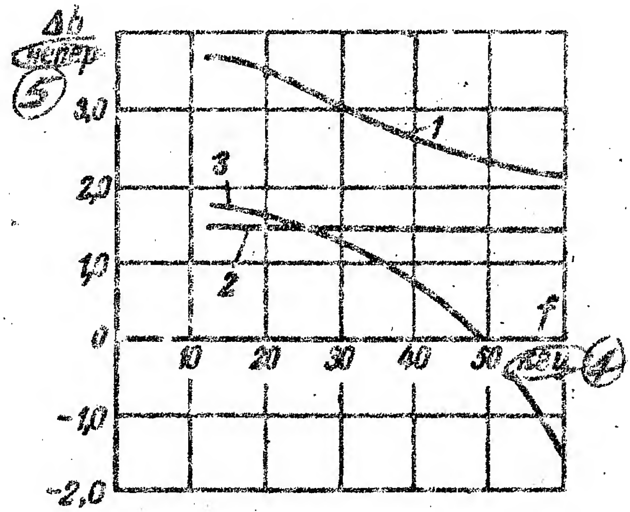


Fig. 4-19. Change in attenuation Δb with a line length of 1000 km and a 10°-C temperature increase.
1--nonloaded line with paper insulation; 2--close-loaded line with styroflex insulation; 3--close-loaded line with paper insulation.

4--kilocycles.
5--nepers

equivalent to the attenuation of paper-insulated cable with a conductor diameter of 1.2 mm.

A major advantage of styroflex cable is the non-dependence of the electrical data on temperature factors.

It will be seen from Fig. 4-19 that when the temper-

ature changes by 10°C, the attenuation of cables with paper insulation flustuates between 1.5 and 3 nepers, while in styroflex-insulated cables the frequency curve of attenuation is constant and rectilinear. This greatly facilitates the adjustment of transmission level and the use of the cable mains.

4-6. Loading Coils

The loading coil is a closed ring-shaped core of circular or oval section wound with insulated copper wire (Fig. 4-20).

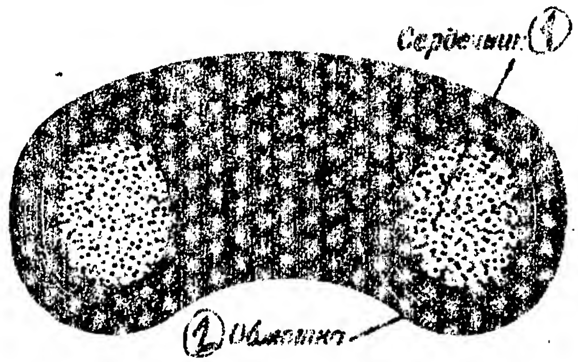


Fig. 4-20. Construction of loading coil.
1) core
2) winding

Loading coils must be characterized by: 1) stability of the specified inductance value and other parameters; 2) minimum active losses ($\tan \varepsilon \rightarrow 0$); 3) minimum

nonlinear distortion; 4) simplicity of construction and minimal dimensions and cost.

To meet these requirements, particularly in the high-frequency region, the cores of loading coils are usually made of a magnetodielectric. The latter is a pressure-formed mixture of ground particles of a ferromagnetic material and a dielectric which serves as a binder. Such ferromagnetic materials as electrolytic or carbonyl iron, nickel alloys, molybdenum permalloy, and alsifer are usually employed, with polystyrol, bakelite resin, shellac, and other high-quality dielectrics as the binders.

The finer the grains of the ferromagnetic powder, the smaller will be the losses and nonlinear distortions in the core. However, reduction of the grain size reduces the magnetic permeability of the material, and it is necessary to choose the appropriate variant in each case in calculating the loading coils.

The basic data of the magnetodielectrics used to fabricate the cores of loading coils are listed in Table 4-10.

The coil winding is made from copper wire 0.6-0.8 mm in diameter, insulated with cotton thread, silk, or viniflex. Wire (litz wire) consisting of a large number of fine insulated wires 0.07-0.1 mm in diameter is used

Table 4-10

(21) Основные параметры магнитодиэлектриков

(22) № по пор.	(23) Наименование материала	μ	(24) $\delta_{вт} \cdot 10^9$ ($f=1$ гц)	(25) $\delta_г \cdot 10^3$ $H=1$ эрст.	(26) $\delta_п \cdot 10^3$
1	Пермаллой ТЧ-180	160—200	1 400	12	2,0
2	„ ТЧ-80	75—85	600	6	1,5
3	„ ВЧ-30	30—33	50	1,8	1,0
4	„ ВЧ-20	17—21	15	1,0	1,0
5	Альсифер ТЧ-60	55—65	200	4,0	1,5
6	„ ВЧ-30	30—33	50	1,8	1,0
7	„ ВЧ-20	17—21	15	1,0	1,0
8	„ ВЧ-8	8—9	25	0,3	0,5
9	„ РЧ-9	9—10	4	0,6	0,6
10	Альсифер повышенной стабильности ТЧ-50	50	200	2,0	1,5
11	Альсифер повышенной стабильности ВЧ-26	26	50	0,9	1,0
12	Карбонильное железо	9—10	3	0,4	0,4
13	Магнетит	6—9	10	1,0	0,5
14	2% Мо-пермаллой	125	380	4	0,6
15	2% Мо-пермаллой	26	32	0,8	0,4
16	2% Мо-пермаллой	14	16	0,36	0,32
17	Карбонильное железо	60	670	5,8	2,9
18	Карбонильное железо	40	11	3,6	1,3
19	Карбонильное железо	12	2,5	0,13	0,16
20	Карбонильное железо	8	1,6	0,05	0,11

1) Permalloy TCh-180
2) " TCh-80
3) " VCh-30
4) " VCh-20
5) Alsifer TCh-60
6) " VCh-30
7) " VCh-20
8) " VCh-8
9) " RCh-9
10) Stabilized alsifer TCh-50
11) " " VCh-26
12) Carbonyl iron [Key continued on Page 156]

13) Magnetite
14) 2% Molybdenum permalloy
15) 2% Molybdenum permalloy
16) 2% Molybdenum permalloy
17), 18), 19), 20) Carbonyl iron
21) BASIC PARAMETERS OF MAGNETODIELECTRICS
22) No.
23) Name of material
24) $\delta_{v.t} \times 10^9$ (f = 1 cps) [eddy-current]
25) $\delta_g \times 10^3$ (H = 1 oersted) [hystersis]
26) $\delta_p \times 10^3$ [aftereffect]

for the windings of high-quality coils.

The loading cables are installed in metallic cases which protect them from mechanical damage and serve simultaneously as electromagnetic screens. The inside of the

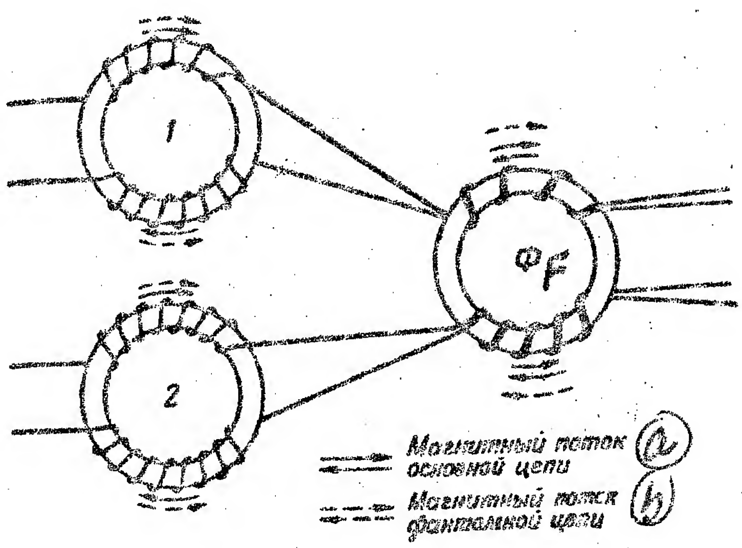


Fig. 4-21. Circuit of loading-coil set. 1--first basic circuit; 2--second basic circuit; F--phantom circuit

a) magnetic flux of basic circuit
b) magnetic flux of phantom circuit

case and the coil which it contains are sealed with a special insulating compound.

In cables with low-frequency loading, where phantom circuits are used, three coils are required for each quad of the cable: two for the basic circuits (1 and 2) and one for the phantom circuit (F). The coil windings are connected in such a way that when current flows through the basic circuits the inductance of the phantom coil is excluded, and when the link is through the phantom circuit the basic coils do not operate (Fig. 4-21).

The loading-coil sets are inserted on a steel frame into a common brass case and then into a protective cast-iron case (Fig. 4-22).

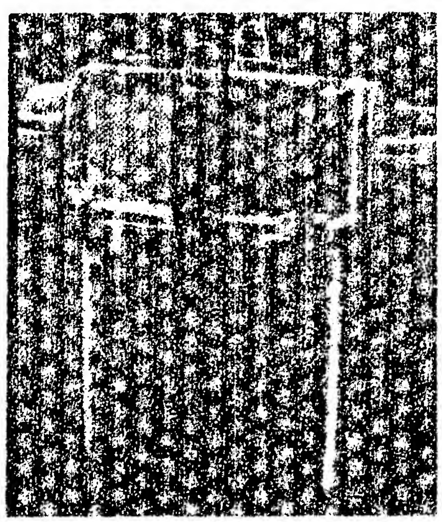


Fig. 4-22. Construction of coil-loading case.

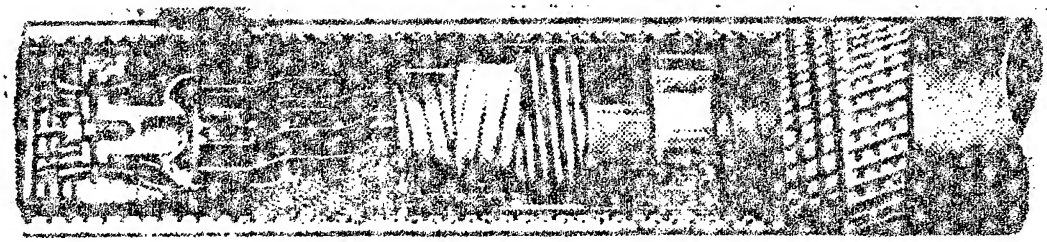


Fig. 4-23. Coil-loading coupling for submarine cable.

The inside of the brass case and the space between it and the iron (outer) case are filled with insulating compound.

Fig. 4-23 shows the placement of the loading coils in flexible couplings used in coil-loaded submarine cables.

Electrical calculations of loading coils are performed as follows:

Starting with a specified value of the inductance L_S, we determine the number of turns N of the winding:

$$N = \sqrt{\frac{L_S D \cdot 10^9}{4 \mu Q}}, \qquad (4\text{-}24)$$

where Q is the sectional area of the core in cm^2;

D is the mean diameter of the coil's core in cm;

μ is the magnetic permeability of the core material.

The (active) resistance of the loading coil consists of the resistance R_0 of the winding to direct cur-

rent and its resistance $R_\sim$ to alternating current, plus the resistance due to eddy-current ($R_{v.t}$) hysteresis (R_g) and aftereffect (R_p) losses in the core:

$$R = R_0 + R_\sim + R_{v.t} + R_g + R_p. \tag{4-25}$$

The value of R_0 is determined by the formula

$$R_0 = \rho \frac{lN}{\frac{\pi d^2}{4}}, \tag{4-26}$$

where ρ is the specific resistance of the winding material in $\frac{\text{ohm-mm}^2}{\text{m}}$;

l is the average turn length in m;

d is the diameter of the winding wire in mm.

The winding's resistance to alternating current

$$R_\sim = \frac{L_S}{\rho\mu} \cdot \frac{V_q}{V_S} f^2 d^2, \tag{4-27}$$

where the volume of the winding $V_q = \frac{\pi d^2}{4} Nl$ [cm^3] and that of the coil $V_S = \pi DQ$ [cm^3].

The hysteresis-loss resistance

$$R_g = \omega L_S \delta_g H, \tag{4-28}$$

where δ_g is the hysteresis loss factor;

H is the magnetic-field strength

$$H = \frac{0.566NI}{D}; \qquad (4\text{-}29)$$

I is the current flowing in the coil.

The eddy-current-loss resistance

$$R_{v.t} = \omega L_S \delta_{v.t} f, \qquad (4\text{-}30)$$

where $\delta_{v.t}$ is the eddy-current loss factor.

The resistance due to magnetic after-effect losses

$$R_p = \omega L_S \delta_p, \qquad (4\text{-}31)$$

where δ_p is the aftereffect loss factor.

The capacitance C for loading coils of various types is given in Table 4-12.

The shunt conductance of loading coils is determined from the formula

$$G_S = \omega C \tan\delta \qquad (4\text{-}32)$$

The value of G_S for the winding-wire insulation of loading coils is exceedingly small and is disregarded altogether in many cases.

The hysteretic properties of the ferromagnetic materials in the cores of loading coils are a cause of nonlinear distortion of the communications signals being

transmitted over the coil-loaded cables. The nonlinear dependence between the current and the voltage leads to the appearance of higher harmonics and combination currents which impair the quality of communication and constitute a source of additional mutual interference between the loaded circuits.

The nonlinear properties of loading coils are characterized by the distortion factor K or the attenuation nonlinearity (klirr factor) B_K.

The values of K and B_K are determined by the hysteresis loss factor δ_g and the magnetic field strength H:

$$K = 0{,}62\delta_z H; \qquad (4\text{-}33)$$

$$B_k = \ln\left|\frac{1}{K}\right| = \ln\left|\frac{1{,}61}{\delta_z H}\right|. \qquad (4\text{-}34)$$

[z = g = hysteresis]

EXAMPLE. Compute the electrical parameters of a loading coil of inductance L_S = 20 millihenries. The loading-coil core has the following data:

$$\mu = 30; \quad \delta_z = 1{,}8\cdot 10^{-3}; \quad \delta_{s.m} = 50\cdot 10^{-9}; \quad \delta_n = 10^{-3}.$$

1) hysteresis; 2) eddy-current; 3) aftereffect

The dimensions of the core are shown in Fig. 4-24

$$(D = 2{,}9\ см;\ l = 5\ см;\ Q = 0{,}85\ см^2).$$

1) cm; 2) cm^2

The winding wire is 0.5 mm in diameter and has viniflex insulation.

The number of turns of the coil

$$N = \sqrt{\frac{L_s D \cdot 10^9}{4\mu Q}} = \sqrt{\frac{20\cdot 10^{-3}\cdot 2{,}9\cdot 10^9}{4\cdot 30\cdot 0{,}85}} = 750 \text{ turns.}$$

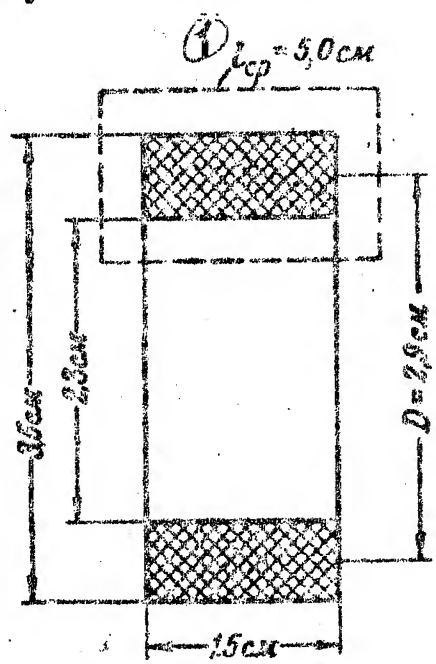


Fig. 4-24. Illustrating calculation of loading coil.

1 $-l_{average}$ = 5.0 cm

СМ = cm

The resistance of the coil

$$R = R_0 + R_{\sim} + R_{z} + R_{d.m} + R_{h};$$

$$R_0 = \rho \frac{lN}{\frac{\pi d^2}{4}} = 0{,}0175 \frac{5\cdot 10^{-2}\cdot 750}{\frac{3{,}14\cdot 0{,}5^2}{4}} = 3{,}4 \text{ ohms};$$

$$R_{\sim} = \frac{L_S}{\rho p}\frac{V_q}{V_S}f^2d^2 = \frac{20\cdot 10^{-3}}{0{,}0175\cdot 30}\cdot\frac{7{,}39}{773}\times$$

$$\times f^2\cdot 0{,}5^2 = 0{,}00454f^2 \text{ ом. (ohms)}$$

where

$$V_q = \frac{\pi d^2}{4}Nl = \frac{3{,}14\cdot 0{,}5^2}{4}\cdot 750\cdot 50 = 7{,}39 \text{ см}^3 \text{ (cm}^3\text{)};$$

$$V_S = \pi DQ = 3{,}14\cdot 2{,}9\cdot 0{,}85 = 773 \text{ см}^3 \text{ (cm}^3\text{)};$$

$$R_{\underset{g}{z}} = \omega L_S\,\delta_{\underset{g}{z}}\,H = 6{,}28f\cdot 20\cdot 10^{-3}\cdot 1{,}8\cdot 10^{-3}\cdot 0{,}584 = 0{,}132f \text{ ом. (ohms)}$$

where

$$H = \frac{0{,}566Nl}{D} = \frac{0{,}566\cdot 750\cdot 4\cdot 10^{-3}}{2{,}9} = 0{,}584;$$

$$R_{\underset{vt}{в.т}} = \omega L_S\,\delta_{\underset{vt}{в.т}}f = 6{,}28f\cdot 20\cdot 10^{-3}\cdot 50\cdot 10^{-9}f = 0{,}0528f^2 \text{ ом; (ohms)}$$

$$R_{\underset{p}{п}} = \omega L_S\,\delta_{\underset{p}{п}} = 6{,}28\;f\,20\cdot 10^{-3}10^{-3} = 0{,}1256f \text{ ом. (ohms)}$$

Note: f is expressed in kc.

The results of calculation of the loss resistance of this loading coil are presented in Table 4-11 for the frequency spectrum to 30,000 cycles.

Table 4-11

(1) **Активное сопротивление пупиновской катушки**

(2) f, кгц	(3) L_S, мгн	(4) R_0, ом	(5) $R_{\sim}$, ом	(6) R_z, ом	(7) $R_{в.т}$, ом	(8) $R_п$, ом	(9) $R_{обмотки}$, ом	(10) $R_{сердечника}$, ом	(11) $R_{общее}$, ом
0,8	20,0	3,4	0,0030	0,106	0,004	0,11	3,403	0,22	3,62
3,0	20,0	3,4	0,0410	0,396	0,076	0,38	3,441	0,83	4,27
10,0	20,0	3,4	0,454	1,32	0,628	1,26	3,854	3,21	7,06
20,0	20,	3,4	1,82	2,64	2,51	2,51	5,22	7,66	12,88
30,0	20,0	3,4	4,10	3,96	5,60	3,77	7,50	13,33	20,83

[Key on next page]

[Key to Table 4-11:]
1) RESISTANCE OF LOADING COIL.
2) f, kilocycles
3) L_s, millihenries
4) R_0, ohms
5) $R_{\sim}$, ohms
6) R_g [hysteresis], ohms
7) $R_{v.t}$ [eddy-current], ohms
8) R_p [after-effect], ohms
9) R_{sheath}, ohms
10) R_{core}, ohms
11) R_{total}, ohms

Table 4-11 seperates the loss resistances in the winding ($R_0 + R_{\sim}$) and the core ($R_g + R_{v.t} + R_p$). The winding losses predominate up to a certain frequency, when they give way to the core losses.

The eddy-current losses increase sharply with increasing frequency. The relative value of the hysteresis losses is comparatively small.

The distortion factor of the coil

$$K = 0{,}62\delta_g H = 0{,}62 \cdot 1{,}8 \cdot 10^{-3} \cdot 0{,}584 = 0{,}654 \cdot 10^{-3};$$

and the klirr factor

$$b_K = \ln \frac{1{,}61}{\delta_g H} = \ln \frac{1{,}61}{1{,}8 \cdot 10^{-3} \cdot 0{,}584} = 7{,}35 \text{ nepers}$$

According to existing standards, the electrical characteristics of the loading coils should conform to

the following data.

1. The inductance, resistance, and capacitance of loading coils should have the values listed in Table 4-12.

Table 4 12

1 Электрические параметры пупиновских катушек

2 Индуктивность катушек, мгн	3 Частота, гц	4 Сопротивление катушек, ом		5 Емкость катушек, пф	
		Основная цепь 6	Фантомная цепь 7	Основная цепь 6	Фантомная цепь 7
140/56	0	8,6	4,3	2 500	1 200
70/29	800	10,1	4,8	2 500	1 200
100/70	1 800	12,8	5,8	2 500	1 200
30/12	0	2,4	1,2	1 200	600
	800	2,7	1,3	1 200	600
	1 800	3,2	1,5	1 200	600
12	0	0,9		600	
	800	1,1		600	
	1 800	1,3		600	
	5 000	2,3		600	
1,75	0	0,7		250	
	30 000	0,9		250	
	60 000	1,3		250	
1,0	0	0,6		200	
	30 000	0,7		200	
	60 000	1,25		200	
3,2	0	1,35		500	
	800	1,45		500	
	[illegible] 000	1,70		500	
	15 000	3,25		500	

[Key to 4-12 on next page]

Key to Table 4-12

1) ELECTRICAL PARAMETERS OF LOADING COILS
2) Coil inductance, millihenries
3) frequency, kc
4) Coil resistance, ohms
5) Coil capacitance, μμfd
6) basic circuit
7) phantom circuit

2. The coil inductance L_S may differ from the rated value by no more than ± 1.5%.

3. At 800 cycles, the unbalance of the half-winding inductances (the percent ratio of the difference in the half-winding inductances to the total) should not exceed:

a) 0.1% for medium- and light-loaded coils;

b) 0.12% for phantom-circuit coils;

c) 0.05% for coils with very light and high-frequency loading.

4. The d-c resistive unbalance of the coil's half-windings should not exceed:

a) 0.1 ohms for medium- and light-loaded coils;

b) 0.03 ohms for coils with very light and high-frequency loading.

5. To limit the nonlinearity introduced by the loading coils, the value of their klirr attenuation B_K should be no smaller than 7.5-8 nepers.

6. The magnitude of the additional resistance arising from hysteresis losses at a frequency of 800 cycles

should not exceed:

a) $12 \sqrt{L_S}$ [ohm/ma-henry] for medium loaded coils;

b) $6 \sqrt{L_S}$ [ohm/ma-henry] for light loaded coils;

c) $2 \sqrt{L_S}$ [ohm/ma-henry] for coils with very light loading.

7. The insulation resistance measured between the winding of one set and all other windings in the case should be no lower than 20,000 megohms.

8. The strength of the dielectric between the windings of one coil should be no lower than 1500 v and that between the winding and the case, 2500 v.

9. Crosstalk attenuation between any circuits of the same set with matched loads over the entire frequency range used should be no lower than:

a) 9.5 nepers for coils with medium, light, and very light loading;

b) 10.5 nepers for radio-broadcasting and loading [sic] coils.

In designing coils with high-frequency loading, it is particularly important to satisfy the requirements with respect to the resistance R and the nonlinear distortion B_K.

It will be seen from Table 4-13, which lists the results of measurement of the values of R_S and tan ε of

loading coils for styroflex-insulated cable (L_S = 1.75 millihenries) that the coil is designed for use in the high frequency region (the minimum tan ε is 2.00 x 10^{-3} at f = 60 kc). Loading-coil losses exert a strong influence on the attenuation of the cable and determine in large part the choice of the coil-loading system of the cable. It is seen from Fig. 4-25 that the optimum coil inductance L_S diminishes with increasing tan ε while the optimum loading interval S increases and the degree of loading of the cable, L = L_S/S, generally declines. The

Table 4-13

Частотная зависимость R_S и tg ε пупиновской катушки индуктивностью 1,75 мгн (1)

(2) f, гц	0	800	5 000	30 000	60 000
(3) R_S, ом	0,66	0,67	0,69	0,91	1,3
(4) tg ε	—	$76,5\cdot10^{-3}$	$12,5\cdot10^{-3}$	$2,76\cdot10^{-3}$	$1,97\cdot10^{-3}$

1) FREQUENCY-DEPENDENCE OF R_S AND tan ε OF LOADING COIL WITH AN INDUCTANCE OF 1.75 MILLIHENRIES
2) f, cycles
3) R_S, ohms
4) tan ε

attenuation of styroflex-insulated cable with a conductor diameter of 1.2 mm, a frequency of 108 kc and a step of

230 m 1s

при $tg\varepsilon = 2{,}65\cdot10^{-3}$ 87 (мнеп/км;
(1) при $tg\varepsilon = 5{,}3\cdot10^{-3}$ 103 (мнеп/км; (2)
при $tg\varepsilon = 10{,}6\cdot10^{-3}$ 125 (мнеп/км.

1) with tan 2) millinepers/km

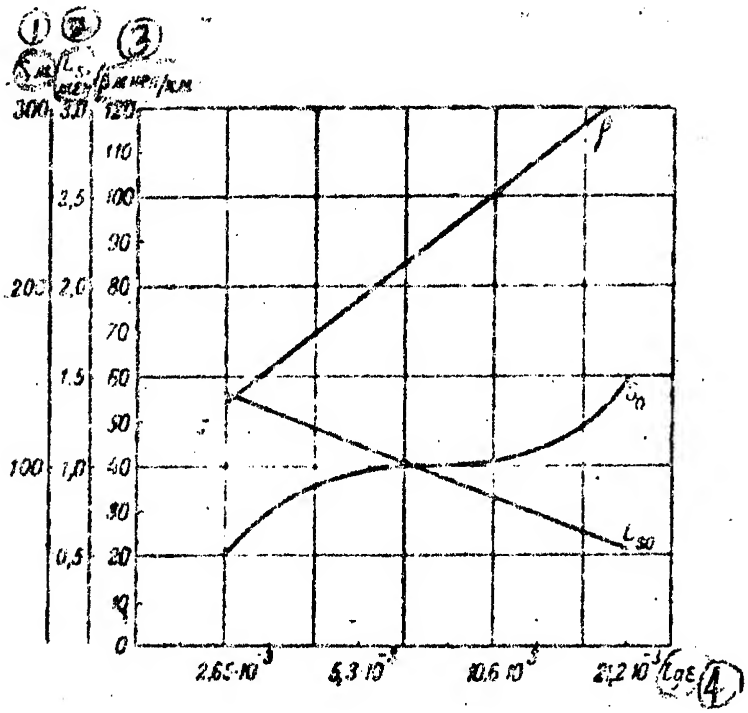


Fig. 4-25. Optimum values of β, L_{s0} and S_0 for different coil losses in styroflex-insulated cable (d = 1.2 mm, f = 108 kc).

1) S, meters
2) L_s, millihenries
3) β, millinepers/km
4) tan ε

4-7. Cables with Ferromagnetic Winding

The inductance of cable circuits may be increased

artificially either by means of loading coils or by winding the conductor of the cable with ferromagnetic tape or wire.

The result of the latter is that a permeable medium is formed about the copper conductor and the magnetic flux and, consequently, the inductance of the cable circuit increase.

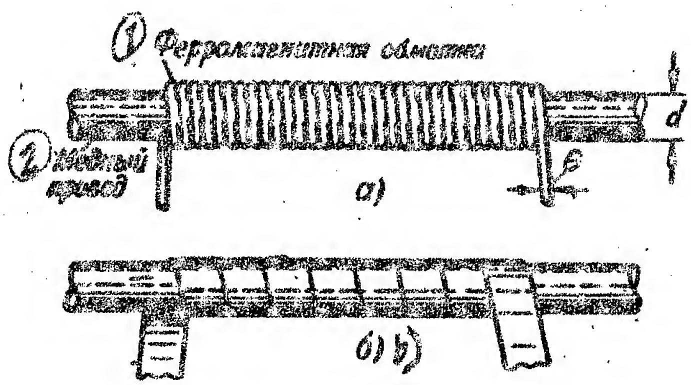


Fig. 4-26. Cable with ferromagnetic winding.

a) normal; wide-strip.

1) Ferromagnetic sheathing; 2) Copper wire

Iron, permalloy (78.5% nickel, 21.5% iron), perminvar (45% nickel, 30% iron, 25% cobalt), and other alloys with high magnetic permeabilities are used as winding materials.

Cables with ferromagnetic windings differ from ordinary communications cables only in the construction of the conductor (Fig. 4 26). The conductor insulation

is gutta-percha, paragutta, paper-packthread, etc.

The "active" resistance R and the inductance L of such cables are computed by the formulas

$$R = R_{м} + \frac{\pi^2 \omega^2 \mu_2^2 \theta^3}{4\rho_2 (d+\theta)} \cdot 10^{-9} \qquad [\text{ohms/km}], \quad (4\text{-}35)$$

$$L = \left(4 \ln \frac{2a-d}{d} + 2\,\frac{\pi \mu_2 \theta}{d} \cos\varphi\right) \cdot 10^{-4} \, [\text{henries/km}], \quad (4\text{-}36)$$

where R_m $[R_{copper}]$ is the d-c resistance of the copper circuit in ohms/km;

θ is the thickness of the ferromagnetic winding in cm;

μ_2 is the permeability of the winding (100-140 for iron);

a is the distance between the conductor centers in cm;

ρ_2 is the specific resistance of the winding;

d is the diameter of the copper conductor in cm;

$\cos\varphi = \dfrac{d+\theta}{\sqrt{(d+\theta)^2+\theta^2}}$ is a correction factor.

Due to the spiral application of the ferromagnetic winding, a bidirectional magnetic field appears about the conductors: a transverse field at right angles to the

axis of the wire, and a longitudinal field which acts along the axis and gives rise to additional losses in the metallic parts of the cable. The effect of the longitudinal magnetic-field component is usually nullified by winding the wire in the same direction around both conductors of the pair or by the use of a two-layered winding with the layers spiraled in different directions. This is the reason for the appearance in (4-36) of the correction factor cos φ, which takes into account only the transverse component of the winding's magnetic field.

The capacitance C, shunt conductance G, and the secondary parameters Z, β, and α of cables with ferromagnetic winding are calculated by the appropriate formulas for symmetrical cables, with the only difference that the conductor diameter includes twice the thickness of the ferromagnetic winding.

It is evident from Expression (4-36) that the inductance L of the circuit is composed of the external and internal inductances determined by the properties of the ferromagnetic winding.

The inductance of the circuit increases as the thickness of the ferromagnetic winding and its magnetic permeability. However, the amount to which the winding can be thickened is limited by the increasing eddy-current

losses in it.

As follows from Formula (4-35), the "active" resistance R of the circuit increases with the thickness of the winding.

For this reason, the wire chosen for winding with the ferromagnetic is no greater than 0.2-0.3 mm in diameter. The use of finer winding wire is difficult from a technological standpoint.

Table 4-14 lists the electrical parameters of cables

Table 4-14

(1) Электрические параметры цепей при толщине обмотки 0,3 мм и частоте 800 гц

(2) d, мм	(3) R, ом/км	(4) L, мгн/км	(5) C, нф/км	(6) G, мкмо/км	(7) β, мнеп/км	(8) α, рад/км	(9) z, ом	φ
1,2	32,7	14	40	0,8	27	0,122	620	12°15′
1,4	24,6	13	42	0,8	22	0,118	570	10°20′
1,5	21,6	12	43	1,0	20	0,115	540	9°55′
1,6	18,7	11,5	44	1,0	18	0,113	520	9°0′
1,8	15,5	10,5	45	1,0	16	0,112	490	8°15′

1) ELECTRICAL PARAMETERS OF CIRCUITS WITH WINDING 0.3 mm THICK AT A FREQUENCY OF 800 kc
2) d, mm
3) R, ohms/km
4) L, millihenries/km
5) C, nanofarads/km
6) G, μmhos/km
7) β, millinepers/km
8) α, rad/km
9) z, ohms

with single-layer ferromagnetic windings of iron wire 0.3 mm in diameter at a frequency of 800 cycles.

The data presented here indicate that the use of a ferromagnetic winding provides for a two- to threefold reduction in the attenuation of the cables.

A major disadvantage of ferromagnetic-wound cables is the narrow frequency band in which they can be used, since the eddy-current energy losses that arise in the ferromagnetic winding with increasing frequency are so great that the latter's advantages are canceled. It will be seen from Formula (4-35) that since ω^2 appears in the numerator of the active-resistance component, which governs the winding losses, $R = \Psi(\omega^2)$ increases as a square-law function. Therefore cables with normal ferromagnetic windings are used primarily for sub-voice-frequency telegraphy (0 to 100 cycles) and voice-frequency telegraphy (300 to 3000 cycles).

Due to their frequency limitations, considerable cost, and difficulty of production, these cables have not been widely used. For practical purposes, they are employed only to cable water obstacles and as inserts in aerial copper circuits. Transatlantic cable communications and telephone-telegraph communication across other oceans are accomplished with the aid of cables whose conductors are wound with ferromagnetic tape.

These cables are considerably more suitable for

laying in water as compared to loaded cables, since there are no boxes with loading coils.

The necessity of expanding transmitted-frequency ranges has led to the appearance in recent years of wide-band cables, which are used for connecting into aerial circuits (inserts; entry into buildings) in the spectrum up to 45 kcps.

The characteristic impedances of such cable and aerial copper circuits are well matched.

The aforementioned expansion of the transmitted-frequency range is obtained by making the ferromagnetic winding from 2-4 mutually-insulated layers of very thin (0.035-mm) tape formed from an alloy of nickel, iron, and magnesium. At a frequency of 45 kc, the inductance of such a cable ($d_{conductor}$ = 1.2 mm) is 13.7 millihenries/km, its capacitance is 35 nanofarads/km, its loss tangent is 0.01, its attenuation 0.2 nepers/km, and its characteristic impedance 650 ohms.

4-8. Cables with Bimetallic Conductors.

As noted previously, an essential shortcoming of coil-loaded cables and cables with ferromagnetic windings is their limited useful frequency range. This is due in the former case to the loading frequency limit and in the

latter to the large losses in the ferromagnetic winding.

Formula (4-35) indicates that the eddy-current losses in the winding may be reduced and the cable's frequency range of utilization expanded accordingly by reducing the thickness of the ferromagnetic winding.

Thus a cable is suitable for use in the spectrum from 50 to 100 kc with a winding thickness θ = 0.03 to 0.05 mm. As noted above, however, the production of cable with so thin a ferromagnetic winding is extremely difficult.

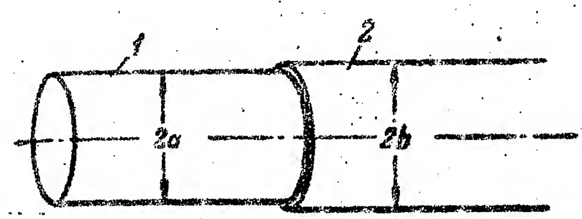


Fig. 4-27. Bimetallic conductor.

In 1938, Prof. I. A. Koshcheyev proposed that the cable conductors be bimetallized by electrical coating with iron (Fig.4-27). The electrolytic method permits the deposition of a thin layer of iron possessing hich magnetic permeability onto the copper conductor; the result is that an increase in inductance is achieved with small eddy-current losses in the ferromagnetic layer over a wide frequency spectrum.

This is illustrated by Fig. 4-28, which shows frequency curves of attenuation for cables with normal and

A) nepers/km

B) kc

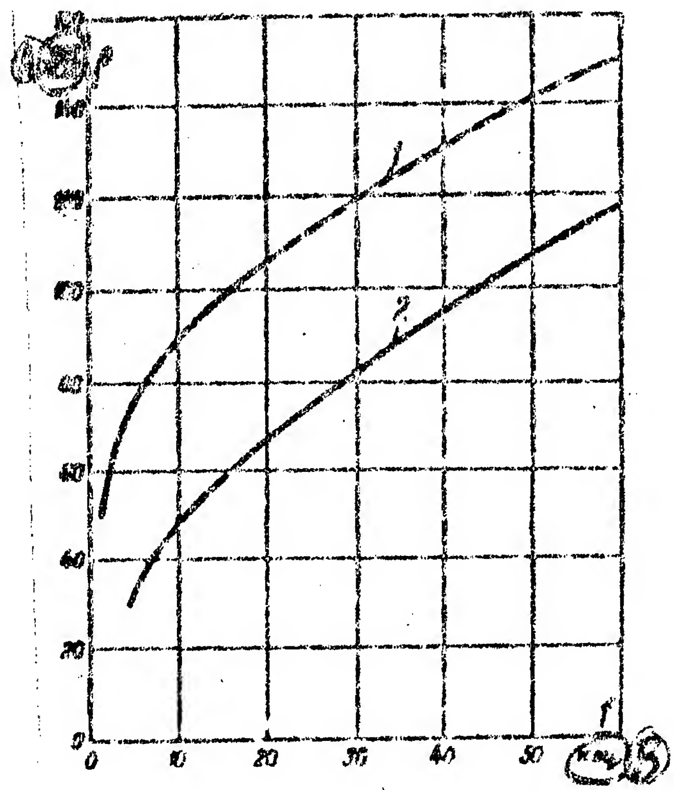


Fig. 4-28. Attenuation of cable with bimetallic conductors as a function of frequency. 1) Copper conductors (d = 1.2 mm); paper insulation; 2) bimetallic conductors (d = 1.2mm, 16-μ layer of iron), paper insulation.

bimetallic conductors, and by Table 4-15.

Table 4-15

(1) Затухание и дальность связи по кабелям с биметаллическими жилами

(2) Наименование характеристик	Кабель с медными жилами, бумажной изоляцией (3)	Кабель с биметаллическими жилами, бумажной изоляцией (4)
(5) Затухание в *мнеп/км* при частоте 108 *кгц*	208	165
Затухание в *мнеп/км* при частоте 60 000 *гц* (6)	150	117
Длина усилительного участка в *км*, при 108 *кгц* (7)	28,8	36,3
Длина усилительного участка в *км*, при 60 *кгц* (8)	40	51

(9) Примечание. Табличные данные получены расчетным путем для кабелей с токопроводящими жилами $d_{жилы} = 1,2$ *мм* при оптимальной толщине железного слоя с $\mu = 100$.

1) ATTENUATION AND COMMUNICATIONS RANGE FOR CABLES WITH BIMETALLIC CONDUCTORS
2) Characteristic
3) Cable with copper conductors and paper insulation
4) Cable with bimetallic conductors and paper insulation
5) Attenuation in millinepers/km at a frequency of 108 kc
6) " " " " " " " " 60,000 cycles
7) Length of repeater section in km at 108 kc
8) " " " " " " " 60 kc
9) Note: the tabulated data were obtained by calculation for cables with conductors having d = 1.2 mm and the optimal thickness of an iron layer having μ = 100.

Table 4-15 presents attenuation figures for cables with ironed conductors and the length of the repeater section with the circuits multiplexed in the frequency spectrum to 60 kc (12-channel systems) and 108 kc (24-channel systems).

It follows from the data given here that the use of bimetallic conductors in paper-insulated cables gives a 21% reduction in attenuation in the range to 108 kc.

The repeater section is 1.4 times longer than in normal cables and amounts to 51 km with 12-channel multiplexing of the circuits and 36 km for the 24-channel multiplexing system.

Even better results can be attained by the use of bimetallic conductors in cables with styroflex insulation.

When ferromagnetics with permeabilities higher than that of iron are used (pentacarbonyl iron, permalloy compositions, etc.), the bimetallic conductor is even more effective. It must be remembered that the optimum sheath thickness is not the same for different ferromagnetic materials, and that the higher the frequency being transmitted through the cable, the thinner must be the ferromagnetic sheath. This is accounted for by the fact that up to a certain frequency, the current flows for the most part through the copper part of the conductor, while with further elevation of the frequency the current expands into the ferromagnetic sheath (due to the so-called skin effect), and the active transmission losses increase. The optimum thickness ϑ of an electrolytic-iron sheath with $\mu = 100$ is tabulated in Table 4-16.

Table 4-16

Оптимальная толщина ферромагнитной оболочки кабел с биметаллическими жилами для различных диапазонс (1) частотного уплотнения кабелей ($d_{жилы} = 1,2$ мм)

(2) Частотный диапа- зон, гц	(3) Магнитная проницаемость	(4) Толщина оболочки (микрон)
30 000	100	25—30
60 000	100	15—18
108 000	100	7—9
156 000	100	4—5

1) OPTIMAL THICKNESS OF FERROMAGNETIC WINDING ON CABLE WITH BIMETALLIC CONDUCTORS FOR VARIOUS RANGES OF FREQUENCY MULTIPLEXING OF CABLES ($d_{conductor} = 1.2$ mm)
2) Frequency range, cycles
3) Magnetic permeability
4) Thickness of sheath, μ

Electrical calculation of the parameters of cables with bimetallic conductors proceeds by the following formulas (all values with the subscript 1 relate to the copper core of the conductor, and those with the subscript 2 to the ferromagnetic sheath):

the "active" resistance of the conductor $R = R_0 x^2 K' \times 10^5$ [ohms/km];

(4-37)

The internal inductance $L_{int} = \frac{\mu_2}{8\pi}K'' \times 10^5$ [henries/km]; (4-38) where

$$x = \frac{b \text{ (the radius of the bimetallic conductor in cm)}}{a \text{ (the radius of the copper part of the conductor in cm)}};$$

$\mu_2 = 4\pi \times 10^{-9}\mu$ is the magnetic permeability of the winding material;

R_0 is the d-c resistance of the bimetallic conductor in ohms/cm;

K' and K'' are coefficients accounting for the influence of the skin effect and expressed in terms of the eddy-current factors $\sqrt{j\omega\mu_2\gamma_2}$ and the radii a and b.

Determination of the coefficients K' and K'' is extremely complex and is accomplished using Bessel functions.

The formulas are vastly simplified for the low-frequency spectrum, i.e., for $\sqrt{j\omega\mu_2\gamma_2}\ a \leq 0.25$,

$$R = \frac{\rho_1}{\pi a^2} \frac{10^5}{1 + \frac{\rho_1}{\rho_2}\left(\frac{b^2}{a^2} - 1\right)} \text{ [ohms/km]}, \qquad (4\text{-}39)$$

$$L_{int} = \frac{\mu_2}{2\pi} \ln \frac{b}{a} 10^5 + 0{,}05 \cdot 10^{-3} \left[\frac{\text{henries}}{\text{km}}\right] \quad (4\text{-}40)$$

where ρ_1 and ρ_2 are the respective specific resistances of

copper and the ferromagnetic sheath in ohm-cm;

$\gamma_2 = \frac{1}{\rho_2}$ is the specific conductance.

Formula (4 10) is valid for the entire frequency spectrum used in practice.

In the high-frequency spectrum, i.e., for $\sqrt{j\omega\mu_2\gamma_2} \times a \gg 5$, R and L_{int} may be calculated from the formulas

$$R = \frac{\sqrt{2}}{\pi b} \cdot \sqrt{\frac{\omega\mu_2}{\gamma_2}}\, 10^5 \left[\mathrm{ohms/km}\right], \qquad (4\text{-}41)$$

$$L_{int} = \frac{\sqrt{2}}{4\pi} \sqrt{\frac{\mu_2}{\omega\gamma_2}}\, 10^5 \left[\frac{\mathrm{henries}}{\mathrm{km}}\right]. \qquad (4\text{-}42)$$

All the expressions given for R and L_{int} apply to one wire, and are doubled in calculating the parameters of two-wire circuits. The over-all circuit inductance consists of the external L_m and the internal L_{int}:

$$L_m + L_{int}. \qquad (4\text{-}43)$$

The remaining parameters are determined from the familiar formulas for symmetrical circuits (see Chapters 2 and 3).

4-9. Cables with Magnetodielectrics

It will be seen from Expression (4-35) that it is possible to reduce the losses in the ferromagnetic sheath of a cable conductor in order to expand the transmitted-frequency spectrum by reducing the sheath thickness θ and increasing the specific resistance ρ. It is not expedient to reduce the magnetic permeability μ, since this is accompanied by a drop in the cable's inductance.

The first course has led to the bimetallic conductor and the second to the creation of cables with magnetodielectrics (see § 4-6).

The foundations of the theory of cables with magnetodielectrics were laid by Candidate of Technical Sciences I. Ye. Yefrimov.

In production, the magnetodielectric layer is applied to the copper conductor of the cable with the aid of an ordinary injection press.

The special nature of the magnetodielectric consists in the fact that while its specific resistance approximates that of a dielectric, it still possesses high magnetic permeability.

As a result, the use of a magnetodielectric sheath increases the inductance of a cable circuit, and due to the large ρ the losses involved remain negligible (the eddy-current loss factor is inversely proportional to the

specific resistance):

$$K = \sqrt{\omega\mu\gamma} = \sqrt{\frac{\omega\mu}{\rho}}.$$

This is illustrated by Fig. 4-29, which shows the frequency curves of attenuation in cables with different specific resistances of the magnetic sheaths.

It will be seen from the diagram that an increase in ρ from 0.17 to 0.82 $\frac{\text{ohm-mm}^2}{\text{m}}$ and a slight modification of the winding's construction reduce the attenuation by a factor of 10, thus permitting a significant expansion of the frequency range transmitted through the cable.

A large number of cable-magnetodielectric compounds exist:

1) An acetone solution of acetylcellulose with powdered permalloy or sendast as a filler;

2) A rubber and polyethylene composition with an alsifer filler;

3) A plastic composition with an iron-nickel powder.

Fig. 4-30 and 4-31 show the dependence of the basic characteristics of a magnetodielectric (magnetic permeability μ and specific resistance ρ) on the ratio of the volumes of the magnetic powder and the dielectric

(the quantity p expresses the percent ratio of the volume of alsifer to the volume of rubber or polyethylene).

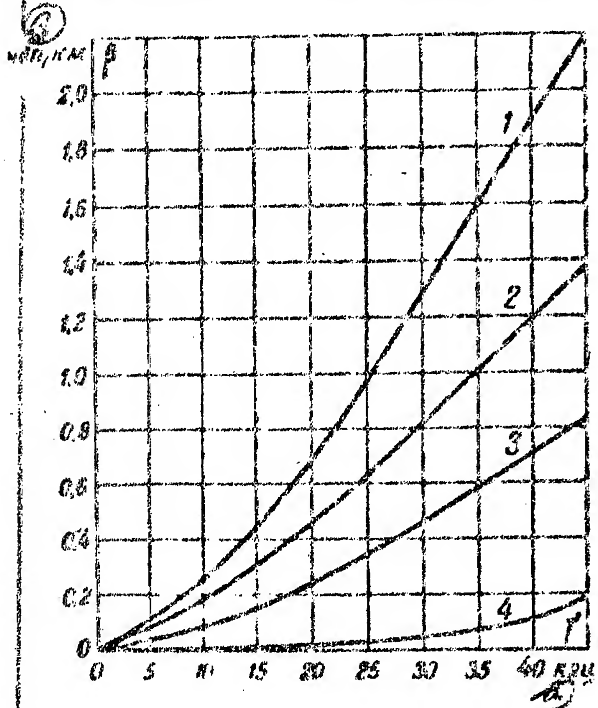


Fig. 4-29. Frequency curve of attenuation of cables with various types of sheathing.

1) ρ = 0.17 ohm-mm^2/m;

2) ρ = 0.3 ohms-mm^2/m;

3) ρ = 0.5 ohm-mm^2/m;

4) ρ = 0.82 ohms-mm^2/m (copper wire with d = 1.5 mm, ferromagnetic winding with δ = 0.3 mm).

A) kc B) nepers/km

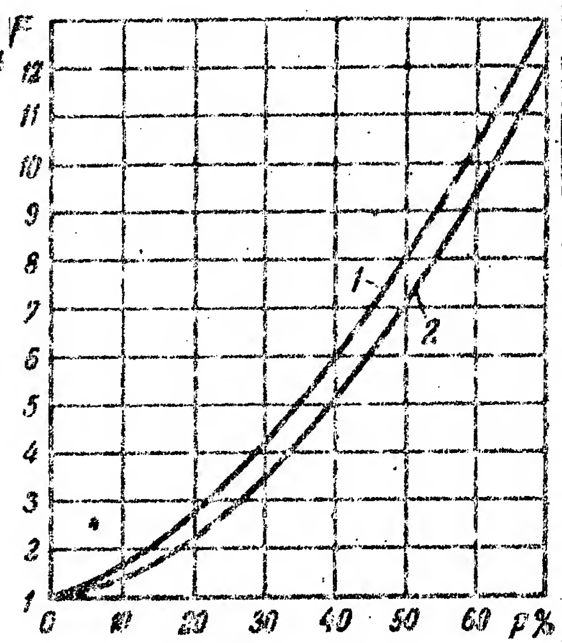


Fig. 4-30. Relative magnetic permeability of magnetodielectrics based on rubber and polyethylene as a function of their degree of saturation with magnetic powder (alsifer).
1) Polyethylene-based magnetodielectric; 2) rubber-based magnetodielectric.

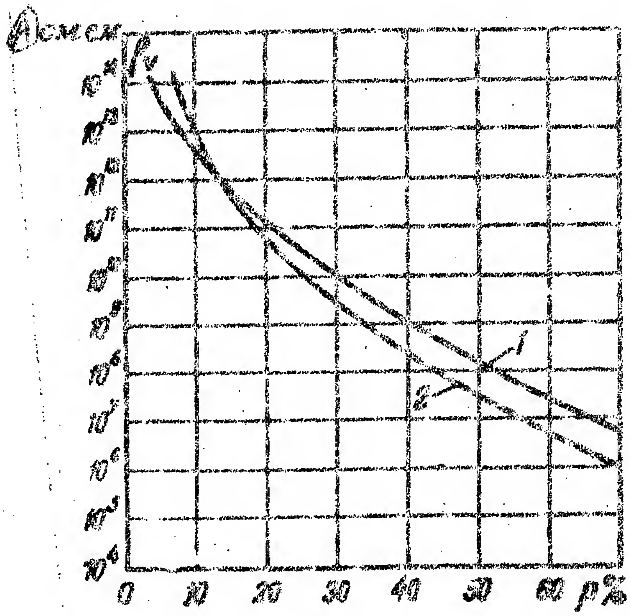


Fig. 4-31. Volume resistivity of rubber- and polyethylene-based magnetodielectrics as a function of filling by magnetic powder.
1) Rubber-based magnetodielectric; 2) Polyethylene-based magnetodielectric.

A) ohm-cm

It will be seen from the diagram that an increase in the degree to which the material is filled with the magnetic powder increases the magnetic permeability of the magnetodielectric with a simultaneous reduction in its specific resistance. The use of compositions with a ferromagnetic content higher than 50-60% is undesirable because it gives rise to a series of technical difficulties in applying the magnetodielectric to the conductor.

The grain size of the alsifer is 50-100 μ.

Given the 50-100-μ grain size in the alsifer and

the above-noted filling, the magnetodielectric has $\mu = 8$ to 10 and $\rho = 10^7$ to 10^8 ohms-cm.

The "active" resistance R and the internal inductance L_{int} of a cable circuit with magnetodielectric is computed from the formulas

$$R = R_0 K' \cdot 10^5 \quad [\text{ohms-km}], \qquad (4\text{-}44)$$

$$L_{int} = \left(\frac{\mu_1}{8\pi} K'' + \frac{\mu_2}{2\pi} \ln \frac{b}{a}\right) \cdot 10^5 \; [\text{henries/km}], \qquad (4\text{-}45)$$

where b is the radius of the conductor with magnetodielectric in cm;

a is the radius of the copper part of the conductor in cm;

μ_1 and μ_2 are the permeabilities of the copper and the magnetodielectric, respectively (a factor of 4 x x 10^{-9} should be taken into account in computing μ_1 and μ_2);

R_0 is the resistance of the copper part of the conductor in ohms/cm;

K' and K" are coefficients taking the skin effect into account. In the low-frequency region, where

$\sqrt{\omega \mu_2 \gamma_2} \times a \leqslant 1$,

$$\left.\begin{aligned} R &= R_0 \cdot 10^5 \quad [\text{ohms/km}], \\ L_{int} &= \frac{\mu_2}{2\pi} \cdot 10^5 \ln \frac{a}{b} \; [\text{henries/km}]. \end{aligned}\right\} \qquad (4\text{-}46)$$

It will be seen from the formulas that the active resistance of conductors with magnetodielectric is equal to the active resistance of the copper part of the conductor. The losses introduced by the magnetodielectric are very small and they may be neglected in practice.

The internal inductance of the wire is equal to the sum of the inductances of the copper conductor and the inductance of the magnetodielectric layer. The latter predominates in magnitude. The values of R and L_{int} found by these formulas should be doubled for calculation of cable-circuit parameters. It is necessary to take the external inductance into account as well in stating the total inductance of the circuit.

The remaining parameters are computed by the general formulas for symmetrical cables.

It should be remembered that the dielectric constant ϵ_m of a magnetodielectric is considerably larger than that of the dielectric itself. Thus, for example, ϵ = 3 to 4 for rubber, but a rubber-based magnetodielectric has ϵ_m = 4.5 to 5. The result is that the capacitance of a cable with magnetodielectric is 30-40% larger than that of an ordinary cable.

The dielectric constant of a magnetodielectric may be determined from the expression

$$\varepsilon_{м} = \frac{\varepsilon \ln \frac{b}{a}}{\ln \frac{b}{\sqrt{pb^2 + a^2(1-p)}}}, \tag{4-47}$$

where p is the filler ratio and ε is the dielectric constant of the original material.

A disadvantage of cable with magnetodielectric as compared with ordinary cable is its considerably greater weight.

CHAPTER FIVE

COAXIAL CABLES

5-1. Special Properties of Coaxial Cables and their Classification

The effort to expand the transmitted-frequency spectrum which is characteristic of the development of communications technology and arose, in this case, from the necessity of creating high-capacity telephone-channel trunks on the most important routes, has resulted, in the last decade, in the use of coaxial cables on such trunk routes. This was also given impetus by the needs of interurban television broadcasting, location, and other special forms of communication. The great and universal interest in coaxial cables is accounted for by the fact that as compared with other types of lines, they come closest to meet-

ing the technical and economic requirements of high-quality communications. The fundamental advantages of coaxial cables are 1) the possibility of transmitting an extremely wide frequency spectrum with relatively small losses, 2) the high degree of protection of the links from the influence of neighboring circuits and external noise, 3) the economy of this communications system as a whole.

We distinguish between trunk- and feeder-type coaxial cables on the basis of their function and design. The former are designed to transmit the frequency spectrum to 8-10 mc, and the latter to transmit up to several thousand megacycles.

The trunk cables are used as long-range interurban communications equipment and transmission of television programs over great distances. They are distinguished by the heavy protective coating which permits their use under all types of conditions (underground, in water, etc.).

The feeder or radio cables are used to connect transmitters with antennas, in the cabling of radio stations, location equipment, and other radio-frequency equipment. A long nomenclature of high-frequency cables of low, medium, and high power, low-capacitance and high-voltage antenna cables, cables with variable characteristic impedance, delay lines, etc. comes under this head-

ing. A distinguishing property of these cables is a high degree of flexibility and elasticity.

The trunk-type radio cables are distinguished by the design of their insulating layer. As a rule, trunk coaxial cables employ composite dielectrics (spacers, packthread, spiral supports, etc.), while radio cables generally use solid insulation.

The basic criterion for the quality of cable (aerial) communications lines is the width of the frequency spectrum which they transmit effectively.

It is natural that the wider this spectrum is, the greater will be the number of different transmissions that can be carried on the cable main in question, and the better will be the technical-economic indices of the system of communication.

The ability to transmit a very wide frequency spectrum is a characteristic property of coaxial designs. In this case, in contrast to that of ordinary symmetrical cables, the high-frequency channels in coaxial cables are in a better position than the low-frequency channels.

Symmetrical cables and aerial lines are suitable for use in a relatively small frequency spectrum. Table 5-1 gives the frequency-utilization spectra of existing wire-communications lines.

Table 5-1

(1) Спектры частот, пропускаемые различными линиями связи

(2) Тип линии	Частотный диапазон, гц (3)	Количество вч связи (4)
Симметричный кабель (5)	60 000	12
Симметричный кабель (6)	108 000	24
Стальная воздушная цепь (7)	10 000	2
Медная воздушная цепь (8)	150 000	15
Пупинизированный кабель с бумажной изоляцией (9)	20 000	3
Пупинизированный кабель со стирофлексной изоляцией (10)	60 000	12
Коаксиальный кабель (11)	3 000 000	660
То же (12)	6 000 000	(13) Телевидение (черно-белое)
» »	10 000 000	(14) Телевидение (цветное)

1) FREQUENCY SPECTRA TRANSMITTED BY VARIOUS COMMUNICATIONS LINES
2) Type of line
3) Frequency range, cycles
4) Number of high-frequency circuits
5) Symmetrical cable
6) Symmetrical cable
7) Steel aerial circuit
8) Copper aerial circuit
9) Loaded cable with paper insulation
10) Loaded cable with styroflex insulation
11) Coaxial cable
12) same
13) Television (black/white)
14) Television (color)

It will be seen from the table that the aerial lines are used in the spectrum to 150,000 cycles, and

this permits the establishment of 15 hf telephone circuits on the line. The basic obstacle to expansion of the frequency spectrum transmitted over copper wires is the increase in mutual interference between the circuits in the channels lying in the upper part of the range. Steel circuits are multiplexed only by 1-2 hf links in the spectrum below 10,000 cycles. The transmission of higher frequencies is limited by the sharply increasing attenuation.

Symmetrical cable circuits are multiplexed by 12 (to 60 kc) or 24 (to 108 kc) telephone links. The transmission of high frequencies over them involves a sharp increase in eddy-current losses in the metallic parts of the cable and, consequently, in increased attenuation. In addition, mutual interference between the circuits rises, and this makes adherence to the normalized crosstalk attenuation difficult.

Due to the presence of the additional-inductance coils, coil-loaded cables are likewise unsuitable for the transmission of high frequencies.

As will be seen from the table, only the coaxial cable permits passage of the frequency spectrum to 3-6 mc which is necessary for transmission of one television program or 60` telephone connections.

5-2. Electrical Processes in Coaxial Circuits

The large frequency-transmission capacity of the coaxial cable is designed into it by concentric arrangement of the forward-transmission conductors inside the return conductor.

The characteristics of propagation of electromagnetic energy along a concentric line open the possibility of subjecting it to broad-band multiplexing and place high-frequency transmission in a favorable position as compared to low-frequency communication. The interaction of the electromagnetic fields of the forward and back wires of a coaxial cable is such that its external field drops to zero.

For simplicity, let us consider the electrical and magnetic fields of the coaxial circuit separately.

The resultant magnetic field of a coaxial cable is shown in Fig. 5-1, which also indicates the magnetic-field strength H_φ^a and H_φ^b for each conductor (a and b) separately.

The magnetic field H_φ^a increases in the metallic interior of conductor a and diminishes outside it according to the law $H_\varphi^a = \frac{I}{2\pi r}$, where r is the distance from the center of the conductor.

The field H_φ^b of conductor b is represented in conformity to the laws of electrodynamics, which establish

that the magnetic field is absent inside the hollow cylinder and is expressed outside it by the same formulas as apply to the solid conductor: $H_\varphi^b = \frac{I}{2\pi r}$, where r is the distance from the center of the hollow conductor. Therefore in determining the external magnetic fields of a coaxial cable the parameter r is assumed identical for the two conductors a and b and reckoned from the center of the conductors (the zero point).

In view of the fact that the currents in the conductors a and b are equal in magnitude and opposite in sign, the magnetic fields of the forward and back conductors (H_φ^a and H_φ^b) will also be equal in magnitude and oppositely directed at any point in space. Consequently, the resultant magnetic field outside the cable will be zero:

$$H_\varphi = H_\varphi^a + H_\varphi^b = \frac{I}{2\pi r} + \left(-\frac{I}{2\pi r}\right) = 0.$$

Thus the lines of force of the coaxial cable's magnetic field are arranged in the form of concentric circles within it. There is no magnetic field outside the cable.

The electric field will also be enclosed within the coaxial circuit and pass in radial directions between the conductors a and b, and will therefore also be zero outside the cable.

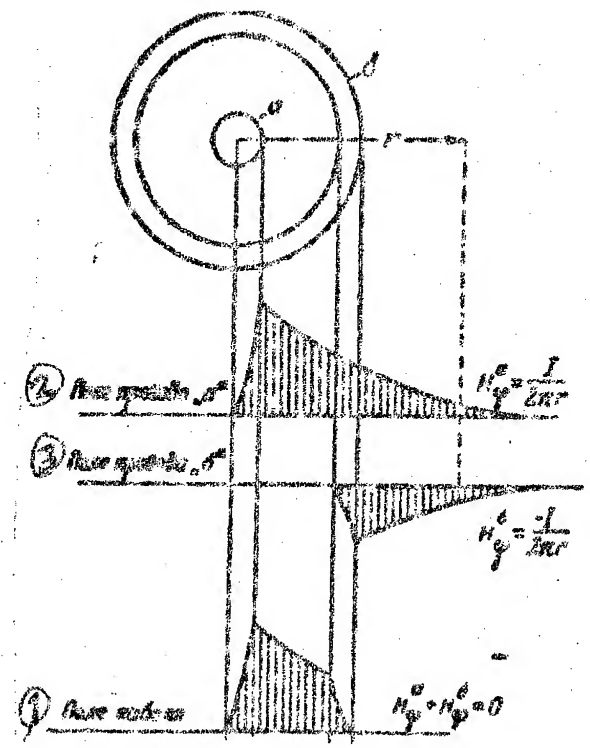


Fig. 5-1. Magnetic field of coaxial cable.
1) field of cable; 2) Field of conductor "a";
3) Field of Conductor "b".

Fig. 5-2 shows the electromagnetic fields of coaxial and symmetrical circuits. It is readily seen that the electromagnetic field of the coaxial circuit is completely contained inside it, while the lines of force of the symmetrical cable's electromagnetic field act at quite considerable distances from it.

The lack of an external magnetic field is responsible for the special merits of coaxial systems.

In ordinary symmetrical circuits, part of the energy is dissipated in the form of thermal eddy-current losses in the neighboring circuits and the metallic masses surrounding the cable (lead sheathing, armor, etc.) due to the presence of the external magnetic field. The external field is lacking in the coaxial cable and no losses of any

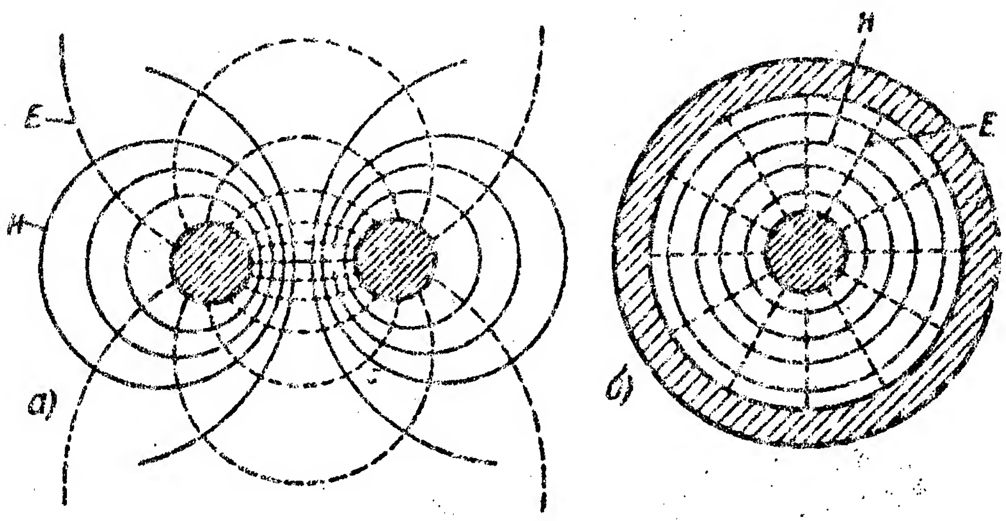


Fig. 5-2. Electromagnetic fields of symmetrical and coaxial cables.
a) symmetrical cable; b) coaxial cable.

kind occur in the metallic components surrounding it. Thus all the energy is propagated inside the cable and transmitted efficiently through it. While the eddy-current losses become so large at a certain point in high-frequency transmission over ordinary cables that communication is impossible, coaxial cables are not host to this deficiency and admit of multiple utilization over a very

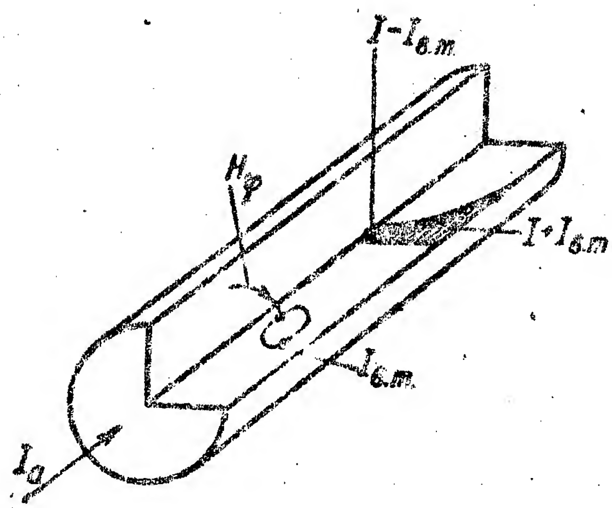


Fig. 5-3. Distribution of current density in inner conductor.

$I_{\delta.m} = I_{v.t}$ = eddy current

wide frequency spectrum. In this connection, it is interesting to consider the influence of the skin effect and proximity effect in coaxial cables and determine the nature of the distribution of current density in the conductors at different frequencies.

The current-density distribution in conductor a is determined only by the skin effect. The effect of the proximity of conductor b to conductor a does not affect the latter, since there is no magnetic field inside the hollow conductor b and it does not influence the current density in the solid conductor a.

The current-density distribution in conductor a is indicated in Fig. 5-3, from which it is evident that an

internal alternating magnetic field that crosses the interior of the conductor will induce therein the eddy currents $I_{v.t}$, which, according to the Lenz law, are directed counter to the corkscrew. Therefore the eddy currents coincide in direction with the basic current at the surface of the conductor ($I + I_{v.t}$) and oppose it in the center of the conductor ($I - I_{v.t}$).

As a result, the current density is significantly higher in zones near the surface than in the central zones of the conductor.

As the frequency of the current transmitted through the circuit increases, the influence of the skin effect becomes stronger and, consequently, the current is drawn increasingly toward the periphery of the conductor.

Redistribution of the current density over the section of the conductor b is governed by the effect of its proximity to the conductor a, since conductor b falls within the sphere of influence of the alternating magnetic field created by the current flowing through the inner (solid) conductor a.

In the absence of conductor a, the current in the hollow conductor b would be drawn toward its periphery by the sk: : effect, as in ordinary solid conductors. However, the proximity effect of conductor a results in a

quite different redistribution of current density in conductor b.

As indicated on Fig. 5-4, the alternating magnetic field set up by the current in conductor a induces eddy currents ($I_{v.t}$) inside the metal of the hollow conductor b; these circulate about the field's lines of force in a direction counter to the rotation of the corkscrew. On the inner surface of the conductor b, the eddy currents coincide in direction with the basic current ($I + I_{v.t}$) and on its outer surface they oppose the latter ($I - I_{v.t}$).

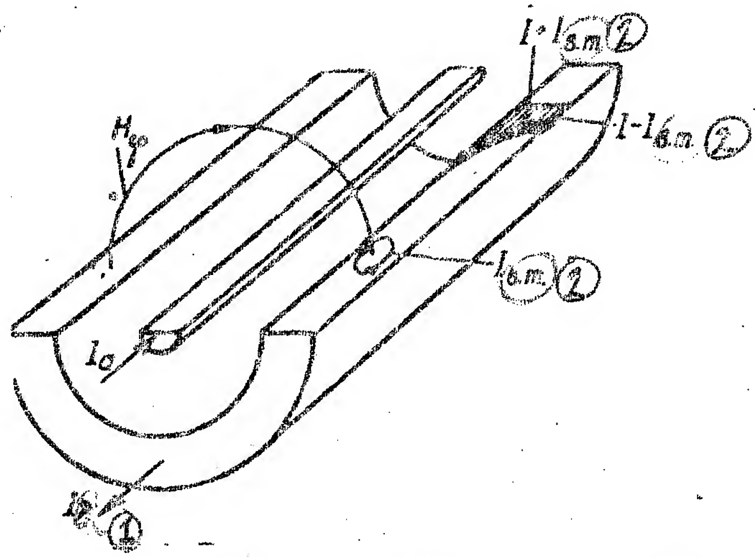


Fig. 5-4. Current-density distribution in outer conductor.

1) b 2) eddy-current

As a result, the current in conductor b is redistributed so that its density increases as we approach the inner surface. Consequently, the currents in conductors a and b will be shifted, and become concentrated on the mutually facing surface of the conductors (Fig. 5-5).

The higher the frequency of the current, the more the lines of current are shifted toward the outer surface of conductor a and the inner surface of conductor b.

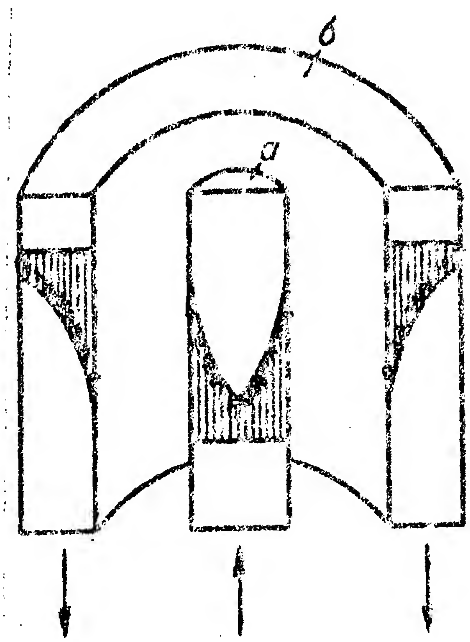


Fig. 5-5. The concentration of current toward the mutually facing surfaces of conductors a and b.

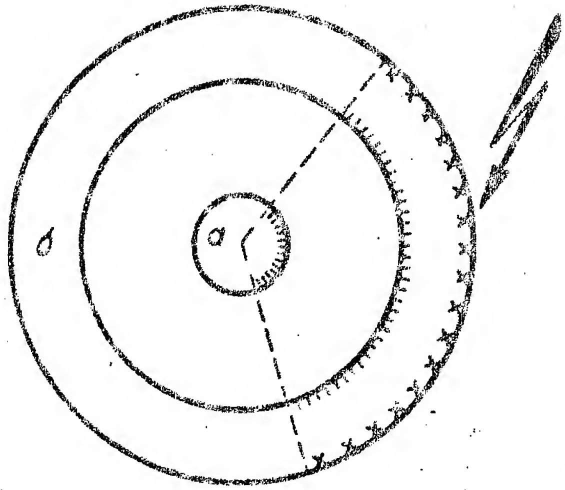


Fig. 5-6. The working current and the noise current in a coaxial cable. A) Working current; B) noise current.

It is as if the energy were displaced from the metal mass of the conductors and concentrated within the coaxial cable in the dielectric, while the conductors serve only to direct the propagation of the electromagnetic waves.

The current-density redistribution effect and the displacement of the current from the surface of the conductors is proportional to the intensity of the eddy currents, and is expressed mathematically by the equation:

$$\sigma = \sqrt{j}\,K = \sqrt{j\omega\mu_1\gamma_1},$$

where K is the eddy-current coefficient.

The interfering high-frequency electromagnetic field,

created by adjacent transmission circuits or other noise sources, and acting at the outer shell (conductor b) of the coaxial line, will also be propagated only around the surface of the cable rather than into its entire cross section. Here, as for the proximity effect, the noise currents will be concentrated at the surface of the conductor that faces the current source. Again, the greater the frequency of the noise current, the less deeply it will penetrate into the outer conductor of the coaxial cable.

Thus, the outer shell of a coaxial cable performs two functions: 1) acts as the return conductor of the transmission circuit (conductor b), 2) protects (screens) the transmission that is preceding along the cable from interfering effects.

It is clear from Fig. 5-6 that the basic transmission current is concentrated at the inner surface of conductor b of the coaxial cable, while the noise current is concentrated on the outside of the outer conductor.

The depth of penetration of both the fundamental current and the noise current into the thickness of the conductor is determined by the eddy-current coefficient. The higher the frequency the greater the distance that separates the 'asic current and the noise current, and, consequently, the better the protection of the cable from the action of extraneous noise.

Thus, in contrast to other types of cable, which require special measures for protection from interference (balancing, screening, etc.), this protection is provided in coaxial cables by the structure itself. Again in contrast to other cable systems, the screening ability of the coaxial cable increases as the frequency rises.

In order to achieve the required degree of protection of the transmission in a coaxial cable from noise, it is necessary to insure that the noise current penetrating into the depth of the outer conductor b does not reach the fundamental current, which is concentrated on its inner surface. Thus, the thickness of the outer conductor is so calculated that there will be a fixed interval between the depth of penetration of noise currents and the basic current.

From what has been said above it follows that the fundamental advantages of a coaxial cable (low attenuation and high resistance to interference) are especially effective in the high-frequency portion of the transmitted frequency band.

With direct current and at low frequencies, where the current occupies practically the entire cross sections of the conductors, the advantages of this cable disappear. Moreover, since a coaxial circuit is unbalanced with res-

pect to other circuits and grounds (the parameters of conductors a and b differ), symmetric cables are better in the low frequency range in all respects.

In accordance with established practice, coaxial cables are used in the 60 kc to 4-10 Mc band, with 60 kc being used as the lower limit for multiplexing of these lines.

This frequency distribution permits coaxial and balanced circuits to be combined effectively into a combination trunk cable with no danger of interaction between them.

5-3. STRUCTURE OF COAXIAL TRUNK CABLES

With respect to dimensional relationships (d/D) the trunk coaxial cables presently in use may be classified as: small, 1.83/6.7, medium, 2.6/9.4, and large, 5/18.

The numerator of the fraction designates the outside diameter of the inner conductor, d, in mm, and the denominator--the inside diameter of the outside conductor, D, in mm.

In many countries, cables are used that have different ratios (3.17/11.7; 2.65/9.55); these may be classified as medium.

In order to become familiar with existing types of

coaxial cables, it is necessary to characterize the remaining structural elements (the inner conductor, insulation, outer conductor, shield, etc.).

I. Inner Conductor

The requirements for this conductor are: a) Good electrical conductivity; b) non-magnetic behavior ($\mu = 1$); c) mechanical strength and adequate flexibility; d) cylindrical shape. There are cables with the following types of inner conductor: Solid, hollow, bimetallic, flexible, twisted from individual thin wires (sometimes enameled).

In the majority of cases coaxial cables are manufactured using a solid copper inner conductor with a diameter ranging from 0.3 to 10 mm. "Protopal" is most often found in the center of bimetallic-type inner conductors. In the 5/18 cable, in order to save copper, the inner conductor is made with an aluminum core coated with copper 0.15 mm thick. Protopal is manufactured by cold working using a machine which in one operation longitudinally notches the aluminum core, applies the copper tape, and rolls it in.

A bimetallic conductor is so designed that the current in it does not penetrate to a depth greater than the thickness of the high-conductivity surrounding shell (copper, silver).

Multi-conductor inner conductors are not found in trunk coaxial cables, since solid conductors have sufficient flexibility.

II. Insulation

The dielectric of a coaxial cable must rigidly preserve the concentricity of the inner and outer conductors both in production and in service. In addition, it must approximate the electrical insulating properties of air as closely as possible ($\tan\delta \approx 0$; $\epsilon \approx 1$; $\rho \approx \infty$), as it is the ideal dielectric. The difficulty of combining these two requirements has led to the creation of rather complicated dielectric structures (beads, supporting spirals, caps, cord frameworks, etc.). Insulation may be subdivided into uniform and combined.

Combination insulation is used exclusively for trunk cables; the ratio of the volumes of dielectric, V_d, and air, V_a, in existing types of cable is approximately $V_d/V_a = (1/10) \div (1/30)$.

Combination insulation is discontinuous or continuous.

A typical variety of discontinuous insulation is bead-type insulation, made by locating beads (20-60 mm) on the inner conductor of a coaxial cable.

Cords, supporting spirals, multi-layer tapes, caps,

and also frameworks are classified as continuous-type insulation.

Dielectrics used for coaxial cables must have:

a) Excellent and stable dielectric properties over a wide frequency band (a low-value of ϵ and tan δ and a high value of ρ and E_{pr});

b) mechanical strength;

c) low hygroscopicity, and properties which do not change over a very long period of service.

Technical treatment of the dielectric must be simple.

Although when coaxial cables were first being produced ceramic, rubberoid and fiber dielectrics were used, at present the basic insulating medium is a plastic, such as styroflex or polyethylene.

Plastics are distinguished by the stability of their excellent electrical characteristics, especially at high frequencies, and by possessing adequate mechanical strength and by the simplicity with which production operations may be performed on them.

Ceramic insulation is chiefly used nowadays for cables intended for high-temperature service.

Production and operating experience has shown that the most acceptable form of insulating structure for trunk

coaxial cables is the bead type of insulation. Its advantages have recently been especially increased in connection with the wide introduction of polyethylene, which permits wide automation of all manufacturing processes involved in the insulation and production of cable.

The physical, mechanical, and electrical properties of various cable dielectrics are considered in detail in Chapter 9.

Here we will consider only the most typical designs and methods for insulating trunk coaxial cables.

a) Bead-Type Insulation

Here the cable is insulated by using beads, spaced at fixed intervals (20-60 mm) along the inner conductor (Fig. 5-7).

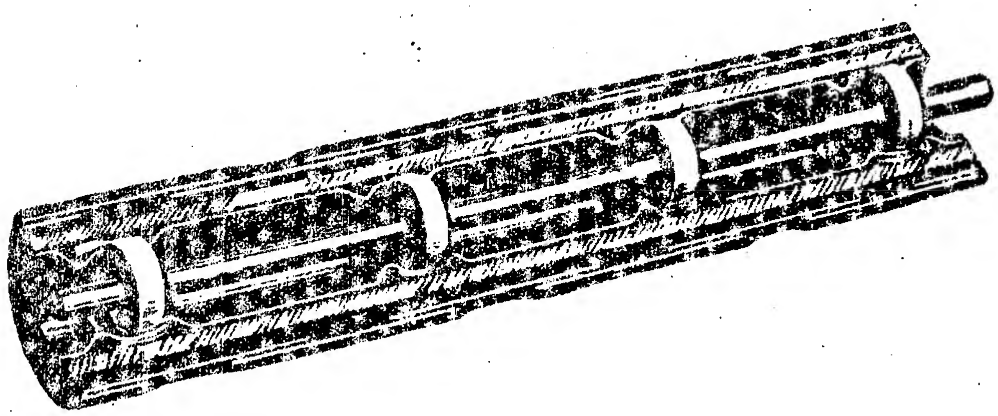


Fig. 5-7. Bead-Insulated Cable

The hole in the bead corresponds to the diameter of the inner conductor; its outside diameter will match the inside diameter of the outer conductor.

The raw material for the beads may be polyethylene, polystyrene, the "frekvent" type ceramic or steatite, as well as solid rubber (ebonite).

Basic design data for insulating beads are given in Table 5-2.

Table 5-2.

DESIGN DATA FOR BEAD-TYPE CABLE INSULATION

Cable Type;	Bead Material;	Bead Spacing, mm;	Thickness of Bead, mm;
5/18	Ceramics;	60	5
5/18	Polystyrene and Ceramic;	60	3
2.6/9.4	Polyethylene;	25	2.2
1.83/6.7	Polyethylene;	20	1.78
1.83/6.7	Ebonite	20	1.6

For installation on the conductor, a lateral notch is cut in polyethylene beads, while in ebonite, polystyrene, or ceramic beads a groove is cut.

In order to simplify the cable structure, the beads are so located that the lateral notches periodically change their position by 180°.

b) Cord-Type Insulation

In this type of insulation, two cords are normally

used; they are wound about the inner conductor in an open spiral (Fig. 5-8). They are used exclusively in small and medium-size cables.

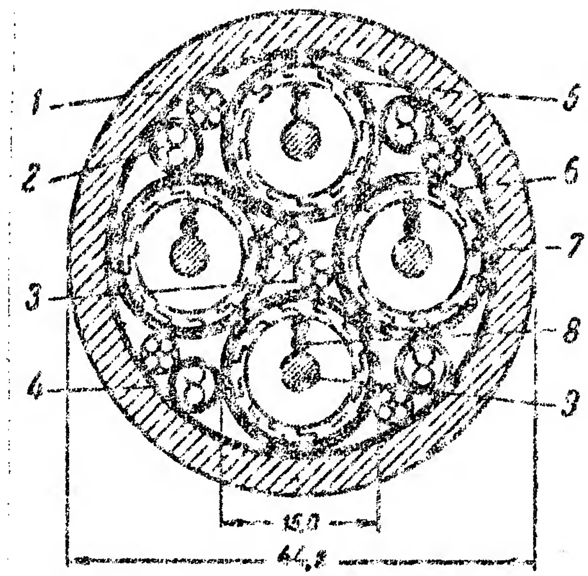


Fig. 5-8. Cord-Insulated Cable With Z-Shaped Outer Conductor
1) lead sheath; 2) symmetric pair; 3) symmetric quad; 4) symmetric-circuit shield; 5) coaxial circuit; 6) coaxial-circuit shield; 7) outer conductor of coaxial circuit; 8) insulation (kotop cord); 9) inner conductor of coaxial circuit.

A drawback to the use of cord-type insulation is the considerable volume of dielectric in the insulating layer, and for this reason the poor electrical properties (ε and tan δ) of the cable.

c) Styroflex Supporting Framework

Structurally, this type of insulation consists of two layers of styroflex combination cord (Fig. 5-9), which takes the form of a spiral with a thin filament running

within it.

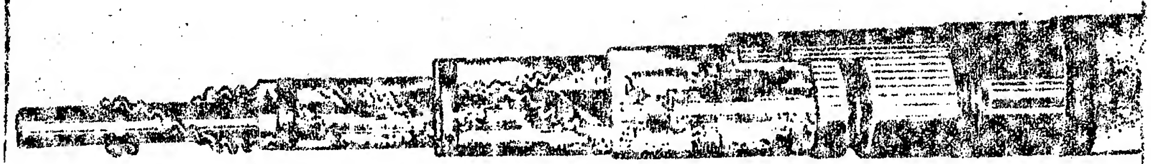


Fig. 5-9. Cable with styroflex frame-type insulation and an outer conductor of overlapping half-tubes.

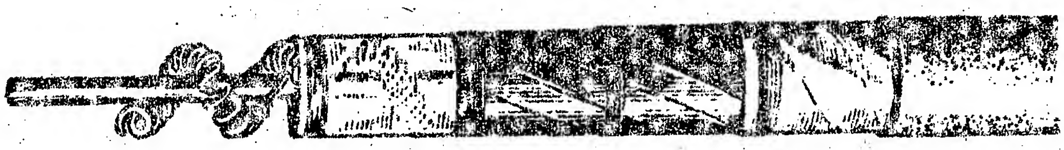


Fig. 5-10. Cable with an outer conductor of ordinary tape.

Over each layer there is a winding of thin styroflex tape. The diameter of the styroflex filament is 0.6-1.4 mm, the thickness of the tape is 0.06-0.1 m.

Frame-type insulation is widely used in large-diameter cables.

Cables having a single layer of spiral-cord styroflex insulation are also encountered (Fig. 5-10).

III. The Outer Conductor

Together with the insulation, one of the most complex structural elements of the cable is the outer conductor.

From the point of view of electrical properties of

the coaxial cable, the best form of outer conductor is a hollow cylinder which is uniform over its entire length. In this case all of the energy is propagated along the cable in the axial direction with no additional losses or distortion.

Punctures, dents, seams, twists, and other non-uniformities in the structure of the outer conductor distort the electrical field within the cable; this results in additional thermal losses owing to eddy currents.

It is exceptionally difficult, however, to manufacture a sufficiently long cable with a cylindrical outer conductor, and in addition, it would not be flexible.

The technical difficulties in producing a cylindrical outer conductor account for the fact that at present there are as many as 15-20 different designs. Of these, the following have found fairly wide use in trunk cables:

a) Conductor of Rectangular Copper Tape

In this type of construction, 12-20 copper tapes, of rectangular cross section (Fig. 5-10), are wound spirally, with large spacing, along the length of the cable, with each tape abutting the next. Above this a thin copper tape is wound in the form of an overlapping lateral spiral, holding the entire structure together.

A substantial drawback to an outer conductor of

this type is the instability of the electrical parameters over long periods of service. The oxide film which forms at the abutting edges of the tapes increase the contact resistance between them and destroy the continuity of the cylinder's conductivity along its perimeter. As a result, the current is forced to travel in a spiral path, which creates a longitudinal magnetic field, and gives rise to additional losses in the transmission circuit.

b) Conductor of Z-Shaped Copper Strips

As shown in Fig. 5-8, 12-24 strips having a Z-shaped cross section are wound in an overlapping spiral with a large longitudinal spacing; this produces adequate continuity of the cylinder's conductivity along the perimeter.

In comparison to the first design, this structure provides more reliable contact between the tapes and accordingly greater stability of the electrical parameters in service.

Until recently, the Z-shaped-strip outer conductor has been widely used in the manufacture of small (1.83/6.7) and medium (2.65/9.55) trunk cables.

c) Tubular Conductor

Tubular conductor (Fig. 5-11) is chiefly used in

large cables having bead-type insulation. It is made of copper or aluminum.

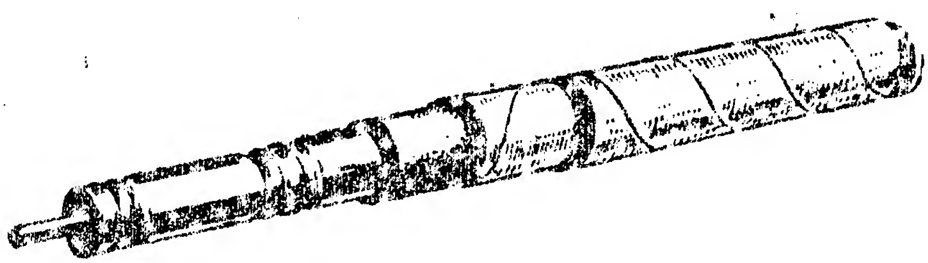


Fig. 5-11. Cable with a tubular outer conductor.

The conductor is made in the form of a continuous single-seam copper tube 0.35 mm thick. Every 50-60 mm, a small longitudinal notch is cut and lateral channels are stamped so as to form a coupling. The insulating beads are located at these points. An outer conductor is also used that consists of 50-70 mm long tubes.

d) Conductor Formed of Half-Tubes

A conductor consisting of a longitudinal copper tube with two longitudinal seams is made by stamping two long copper strips 0.35 mm thick into semi-cylinders (Fig. 5-9). Lateral channels are made every 20 mm along the entire length of the strips. In order to strengthen the two-seamed cylinder, the strips are assembled with the lateral channels displaced with respect to one another.

Both the single-and-double seamed tubular conductor are used in large cables.

e) "Zipper"-Type Conductor

This is conductor made of a continuous cylindrical tube having a single longitudinal zipper-seam (Fig. 5-12). A 0.25 mm thick copper strip is used.

As is shown in Fig. 5-12, the teeth on the edges of the strip are offset, and when the strip is bent, a rigid and quite durable cylinder is formed.

Structurally and electrically this type of outer conductor is the nearest to an ideal cylinder. One of its faults is insufficient flexibility.

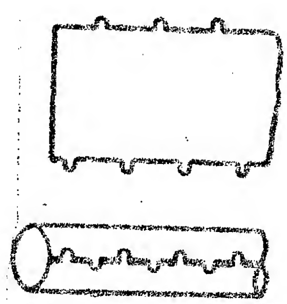


Fig. 5-12. "Zipper"-Type Outer Conductor

This conductor design was developed comparatively recently, and is widely employed in the manufacture of small and medium-sized cables. It is too rigid to be used easily in large cables.

IV. SHIELDING AND CABLE ARMORING

For protection against internal and external noise at low frequencies (60 kc), coaxial cables are generally provided with shielding. Shielding is also necessary in order to screen out the external electromagnetic field which is caused by eccentricity of the inner and outer conductors, which is unavoidable in production. At the same time, shielding increases the mechanical strength of the cable and contributes to the stability of its characteristics.

As a rule, shielding consists of a winding of two steel tapes having thickness of 0.15-0.30 mm and width of 10-15 mm, wound around the outer conductor of the cable so as to overlap.

A shielding shell is chiefly used in single-cable communications systems, where several coaxial pairs are located in the same cable.

The balance of the sheathing (lead and armoring) of coaxial cables is the same as that used in symmetric cables.

5-4. The Eddy-Current Coefficient and the Depth to Which an Alternating Current Penetrates into the Conductors

In order to make a quantitative evaluation of the

electrical processes occurring in coaxial cables, and to compute their parameters, it is first necessary to consider the expressions for the eddy coefficient K, and for the equivalent depth to which an alternating current penetrates into the metal, θ.

It has been shown above that eddy currents force the basic currents toward the surface of the conductor that faces the source of the back current. The effect of eddy currents is proportional to the frequency of the transmitted current, as well as to the electrical conductivity and magnetic permeability of the conductor metal.

The eddy currents are expressed in terms of the coefficient

$$\sigma = \sqrt{j}K = \sqrt{j\omega\mu_1\gamma_1} = \frac{K}{\sqrt{2}} + j\frac{K}{\sqrt{2}} = Ke^{j45^\circ}, \qquad (5\text{-}1)$$

where the coefficient K characterizes the attenuation of energy within the metal, while the angle represents the phase shift of the current as it passes along the metal.

Unlike direct current, alternating current does not flow uniformly through the entire cross section of a conductor, but concentrates to a fixed radial depth along the entire periphery of the conductor. Thus the effect of eddy currents and the surface effect caused by them may

be expressed in terms of the equivalent depth θ to which the currents penetrate into the conductor. This depth is numerically defined to be the thickness of the wall of a hollowed cylindrical conductor having a direct-current resistance equal to the resistance of a solid conductor of the same diameter to alternating current (Fig. 5-13) (R_0 of the hollow cylinder = $R_{\sim}$ of the solid cylinder).

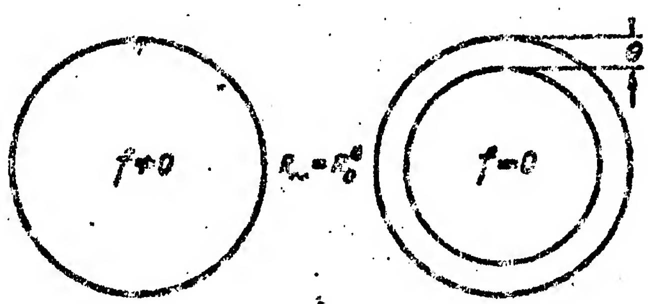


Fig. 5-13. The equivalent depth of penetration of an alternating current into a conductor (θ).

$R_{\sim}$ is the resistance of a solid cylinder to alternating current; R_0^{θ} is the resistance of a hollow cylinder to direct current.

It is clear that the greater the frequency and the stronger the surface effect, the thinner the wall of the equivalent hollow cylinder, and correspondingly the greater the resistance of the conductor.

The stronger the eddy currents, the shorter the distance to which currents penetrate into the metal; the distance is inversely proportional to the real component

of the eddy-current coefficient.

$$\theta = 1 : \frac{K}{\sqrt{2}} = \frac{\sqrt{2}}{K} = \sqrt{\frac{2}{\omega\mu_1\gamma_1}}. \tag{5-2}$$

To simplify practical computations, formulas (5-1) and (5-2) are used in a somewhat revised form:

In practical units,

$$\mu_1 = \mu \cdot 4\pi \cdot 10^{-9},$$
$$\gamma_1 \left[\frac{Mho}{CM}\right] = 10^4 \gamma \left[\frac{Mho\ M}{MM^2}\right].$$

then

$$K = \sqrt{\omega\mu_1\gamma_1} = 2\sqrt{2}\pi\sqrt{\mu\gamma f \cdot 10^{-5}} = 8{,}85\sqrt{\mu\gamma f \cdot 10^{-5}} \tag{5-3}$$

and accordingly

$$\theta = \frac{\sqrt{2}}{K} = \frac{1}{2\pi}\sqrt{\frac{10^5}{\mu\gamma f}}. \tag{5-4}$$

Table 5-3 gives the values of μ and γ in practical units for various metals. The same table gives the expression to be used in calculating the eddy-current coefficient K and θ, the depth to which an alternating current penetrates into the metal.

TABLE 5-3.

The Eddy-Current K and the Depth of Penetration, θ, for Various Metals.

(A) Наименование металла	(F) ρ_s (ом мм²/м)	(G) γ_s (мом/мм²)	μ	K, 1/см	θ, см
Медь (B)	0,0175	57	1	$0{,}21\sqrt{f}$	$6{,}7/\sqrt{f}$
Алюминий (C)	0,0291	34,36	1	$0{,}164\sqrt{f}$	$8{,}6/\sqrt{f}$
Сталь (D)	0,139	7,23	100	$0{,}75\sqrt{f}$	$1{,}88/\sqrt{f}$
Свинец (E)	0,221	4,52	1	$0{,}059\sqrt{f}$	$24{,}0/\sqrt{f}$

A) Metal; B) Copper; C) Aluminum; D) Steel; E) Lead; F) Ohms·mm²/m; G) mho.m/mm².

Table 5-4 and Table 5-14 give θ as a function of frequency for various metals: They show that as the frequency of the transmitted current increases, the depth of penetration sharply decreases. Thus, for copper the depth of penetration at $f = 10^3$ cps is 0.21 cm, while at $f = 10^6$ cps it is only 0.0067 cm.

In a steel conductor, the eddy-current effect is stronger than in copper, and thus θ is 3.6 times smaller for steel than for copper.

In comparison to other cable metals, lead allows the greatest depth of penetration of a current.

TABLE 5-4.

f, гц A	B Глубина проникновения, см			
	C Медь	D Алюминий	E Сталь	F Свинец
10^3	0,21	0,27	0,06	0,76
$6 \cdot 10^4$	0,0272	0,035	0,0077	0,098
10^5	0,021	0,027	0,006	0,076
10^6	0,0067	0,0086	0,0019	0,024
10^7	0,0021	0,0027	0,0006	0,0076
10^8	0,00067	0,00086	0,00019	0,0024
10^9	0,00021	0,00027	0,00006	0,00076
10^{10}	0,000067	0,000086	0,000019	0,00024

Depth of Penetration of Current into Various Metals as a Function of Frequency. A) f, cps; B) Depth of Penetration, cm; C) Copper; D) Aluminum; E) Steel; F) Lead.

5-5. Calculating Coaxial-Cable Parameters

At frequencies of 60 kc and above, the coaxial-cable parameters R, L, C, and G may be computed according to the following formulas.

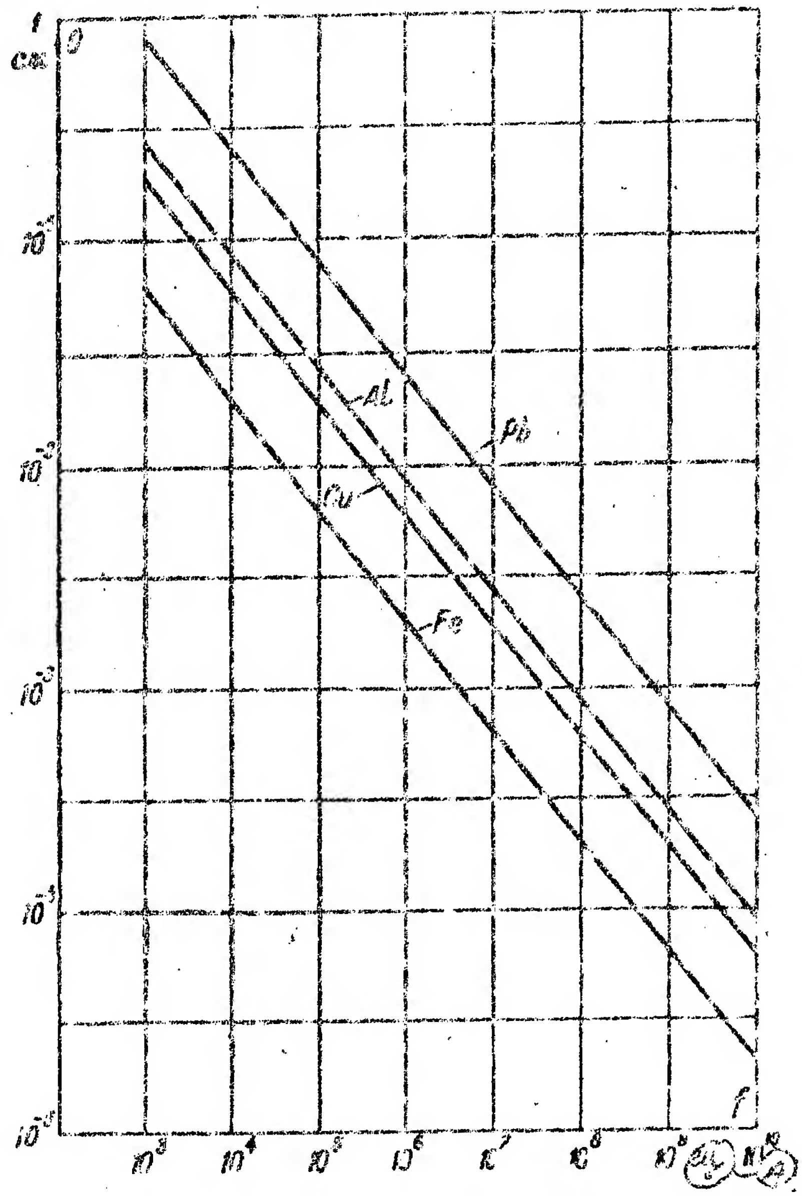


Fig. 5-14. Depth of Penetration of Current for Various Metals. A) cps.

1. The resistance R of a coaxial circuit is directly proportional to the eddy-current coefficient K = $\sqrt{\omega\mu_1\gamma_1}$, and inversely proportional to the diameters d and D and the conductivity of the conductor metal, γ.

The resistance is composed of the resistance R_d of the inner conductor and R_D, the resistance of the outer (hollow) cylindrical conductor.

$$R = R_d + R_D = \left(\frac{K}{2\sqrt{2\pi\gamma_1 r_1}} + \frac{K}{2\sqrt{2\pi\gamma_1 r_2}}\right)[\text{ohm}/\text{cm}], \quad (5\text{-}5)$$

where

$$r_1 = \frac{d}{2}\ (\text{cm}), \quad r_2 = \frac{D}{2}, \quad \gamma_1 = \gamma \cdot 10^4$$

or in the units commonly used in communications work (expressed per kilometer) we obtain:

$$R = R_d + R_D = \left(\frac{K \cdot 10^3}{\sqrt{2\pi\gamma d}} + \frac{K \cdot 10^3}{\sqrt{2\pi\gamma D}}\right)[\text{ohm}/\text{km}], \quad (5\text{-}6)$$

where $K = \sqrt{\omega\mu_1\gamma_1} = 2\sqrt{2\pi}\sqrt{f\mu\gamma \cdot 10^{-5}}$;

γ is the conductivity in mho·m/mm^2 (e.g., for copper $\gamma = 57$);

D and d are the diameters of the outer and the inner conductors in mm;

μ is the magnetic permeability (e.g., for copper, $\mu = 1$).

By using the numerical value of the eddy-current coefficient in formula (5-6), we obtain an expression

which is simpler for computation

$$R = R_d + R_D = \left(\sqrt{\frac{\mu f}{\gamma}} \cdot \frac{2}{\sqrt{10}} \cdot \frac{1}{d} + \sqrt{\frac{\mu f}{\gamma}} \cdot \frac{2}{10} \cdot \frac{1}{D}\right) [\text{ohm}/\text{км}]. \quad (5\text{-}7)$$

If both conductors are made from the same metal, formula (5-7) takes the form:

$$R = \sqrt{\frac{\mu f}{\gamma}} \cdot \frac{2}{\sqrt{10}} \left(\frac{1}{d} + \frac{1}{D}\right) [\text{ohm}/\text{км}]. \quad (5\text{-}8)$$

The values of $\sqrt{\frac{\mu f}{\gamma}}$ are given below for various metals. A) Metal; B) Copper; C) Aluminum; D) Steel; E) Lead:

(A) Наименование металла	$\sqrt{\frac{\mu f}{\gamma}}$
(B) Медь	$0{,}132\sqrt{f}$
(C) Алюминий	$0{,}171\sqrt{f}$
(D) Сталь	$37{,}2\sqrt{f}$
(E) Свинец	$4{,}7\sqrt{f}$

For a coaxial cable with copper conductors, formula (5-8) may be written:

$$R = R_d + R_D = 0{,}0835\sqrt{f}\left(\frac{1}{d} + \frac{1}{D}\right) [\text{ohm}/\text{км}]. \quad (5\text{-}9)$$

where d and D are the diameters of the conductors in millimeters.

Table 5-5 and Fig. 5-15 and 5-16 give the results of the computation of the resistance R and other parameters for a type 2.6/9.4 coaxial cable for the 7-10 Mc frequency band. It is clear from the frequency chart that the increase in the resistance is determined by the value of $\sqrt{f}$.

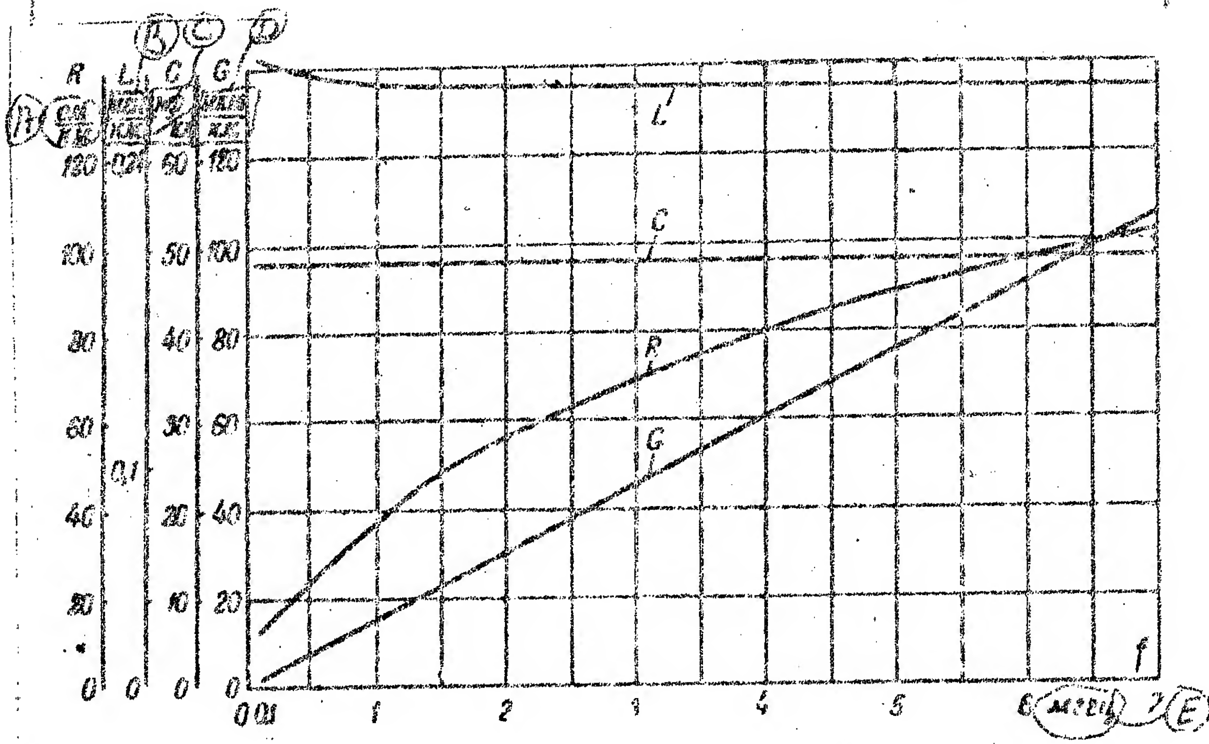


Fig. 5-15. Frequency Dependence of the Primary Parameters for a 2.6/9.4 Coaxial Cable. A) Ohm/km; B) Mh/km; C) Millimicrofarads/km; D) Micro mho/km; E) Mc.

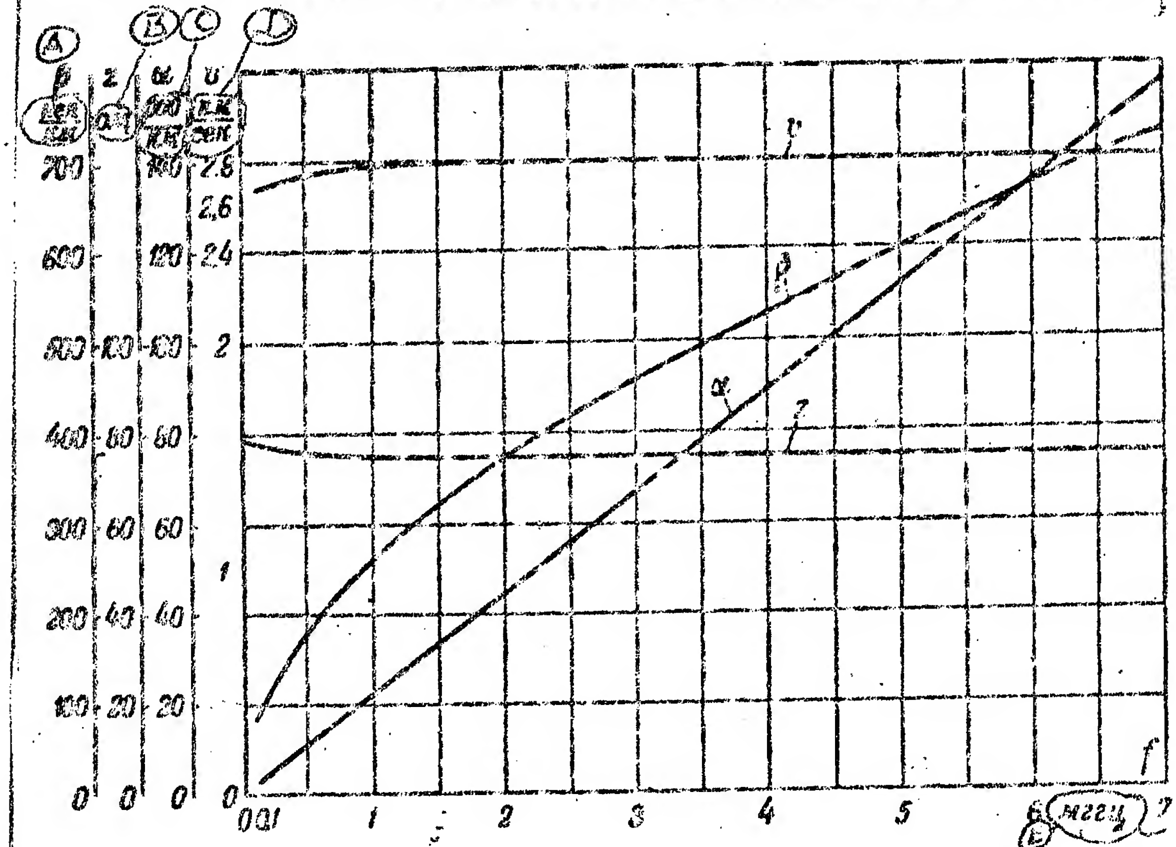


Fig. 5-16. Frequency Dependence of the Secondary Parameters for a 2.6/9.4 Coaxial Cable. A) Neper/km; B) Ohms; C) Rad/km; D) Km/sec; E) Mc.

As the diameters d and D of the conductors increase, their resistance decreases; the inner conductor has the greatest resistance R.

For the dimensional relationships (D/d = 3.6), used in existing cables, $R_d/R_D = 3.6$, i.e., the inner conductor has about 80% of the resistance, and the outer conductor only about 20%.

TABLE 5-5.

(A) f, гц	(B) R_d, ом/км	(C) R_D, ом/км	(D) R, ом/км	(E) $L_{мн}$, мгн/км	(F) $L_{вн}$, мгн/км	(G) L, мгн/км	(H) C, нф/км	(I) G, мкмо/км	(J) α, мнеп/км	(K) β, рад/км	(L) v, км/сек	(M) z, ом
$6\cdot10^4$	8,4	2,79	11,19	0,266	0,022	0,288	48	0,91	72	1,39	$2{,}7\cdot10^5$	77,5
10^5	9,62	2,81	12,45	0,266	0,017	0,283	48	1,52	81	2,32	$2{,}71\cdot10^5$	76,8
10^6	30,5	8,9	39,4	0,266	0,007	0,273	48	15,2	260	22,7	$2{,}77\cdot10^5$	75,3
$3\cdot10^6$	52,8	15,4	68,2	0,266	0,004	0,270	48	45,6	457	68	$2{,}78\cdot10^5$	75,0
$7\cdot10^6$	80,4	22,6	103	0,266	0,002	0,268	48	106	730	157	$2{,}78\cdot10^5$	74,5
10^7	96,4	28,1	124,5	0,266	0,002	0,268	48	152	840	225	$2{,}78\cdot10^5$	74,5

Electrical Parameters of Trunk Coaxial Cable (2.6/9.4) having Polyethylene Insulation. A) Frequency, cps; B) R_d, ohms/km; C) R_D, ohms/km; D) R, ohms/km; E) L_{ic}, mh/km; F) L_i, mh/km; G) L, mh/km; H) C, μμf/km; I) G, μmho/km; J) β, m neper/km; K) α, rad/km; L) V, km/sec; M) Z, ohm.

2. The circuit inductance is composed of the internal inductance of the conductors $L_i = L_d = L_D$ and the inter-conductor inductance L_{ic}:

$$L = L_{ic} + L_d + L_D = \left(2\ln\frac{D}{d} + \frac{20\sqrt{2\mu}}{Kd} + \frac{2\sqrt{2\mu}}{KD}\right)\cdot 10^{-4} \quad \text{[Henrys/km]}, \tag{5-10}$$

where D and d are given in mm,

$$K = 2\sqrt{2}\pi\sqrt{f\mu\gamma\cdot 10^{-8}}.$$

Formula (5-10) is considerably simplified for cables with copper conductors

$$L = \left[2\ln\frac{D}{d} + \frac{133{,}3}{\sqrt{f}}\left(\frac{1}{d} + \frac{1}{D}\right)\right]\cdot 10^{-4}\ [\text{Henrys}/\text{KM}] \quad (5\text{-}11)$$

It follows from formula (5-10) and Table 5-5 that the internal inductance of the conductors drops sharply as the frequency increases. Thus, at a frequency of $6\cdot 10^4$ cps, L_i constitutes 7-8% of the overall inductance, while in the higher frequency region its relative value is still less and does not exceed 0.8%.

In view of this fact, the inductance of a coaxial circuit may be computed with a fair degree of accuracy from the formula

$$L = L^{a} = 2\ln\frac{D}{d}\cdot 10^{-4}\ [\text{Henrys}/\text{KM}] \quad (5\text{-}12)$$

The inductance of existing types of trunk cables is 0.26-0.27 mh/km.

3. The capacitance of a coaxial cable is determined from the formula for a cylindrical capacitor:

$$C = \frac{\varepsilon}{18 \ln \frac{D}{d}} \; [\mu f / \text{km}]. \tag{5-13}$$

from which it follows that the capacitance depends upon the relationship of the dimensions of the cable conductors and the equivalent dielectric constant ε. As D/d increases, the capacitance decreases, a fact which is widely used in the design of low capacitance coaxial cables.

One of the most efficient ways to decrease the capacitance of the cable is to decrease the parameter ε.

It should be noted, that for the same relationships of the diameter of the conductors, the capacitance of a cable having a solid uniform dielectric (ε = 2.3) is 100 $m\mu f$/km, while if combined air-bead insulation (ε = 1.1) is used, it drops to 48-50 $m\mu f$/km.

4. <u>The conductance of the insulation</u> increases in direct proportion to the frequency of the current and is basically determined by the dielectric loss factor tan δ.

$$G = \omega C \tan \delta \; [\text{mho} / \text{km}]. \tag{5-14}$$

When poor dielectrics lead to a large tan δ, the value of G is impermissibly large. In a cable having polyethylene insulation, the shunt conductance at $7 \cdot 10^6$

cps is 106 μmho/km.

5. Since in practice coaxial cables are used at 60,000 cps and above, where R is much less than ωL and G is much less than ωC, their secondary parameters may be computed from the following formulas*: The attenuation constant is

$$\beta = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}}\,[\text{nepers/km}]$$

the phase constant is $\alpha = \omega\sqrt{LC}\ [rad/km]$;

the wave impedance is $z = \sqrt{\frac{L}{C}}\ [ohms]$.

*See Chapter II of this book.

It is desirable, however, to express the secondary parameters of coaxial cables directly in terms of the dimensional relationships within the coaxial cable.

a. Wave Impedance. Using formulas (5-12) and (5-13), we obtain an expression for the wave impedance:

$$z = \sqrt{\frac{L}{C}} = \frac{60}{\sqrt{\varepsilon}}\ln\frac{D}{d}. \tag{5-15}$$

It follows from Table 5-5 that the wave impedance

z changes very little, and it may be considered to be constant.

The dependence of z on ε, given in Fig. 5-17, shows that in a cable with a solid dielectric ($\varepsilon \approx 2.3$), $z \approx 50$ ohms, while in cables with combination insulation ($\varepsilon \approx 1.1$) the value of the wave impedance ranges from 70-75 ohms.

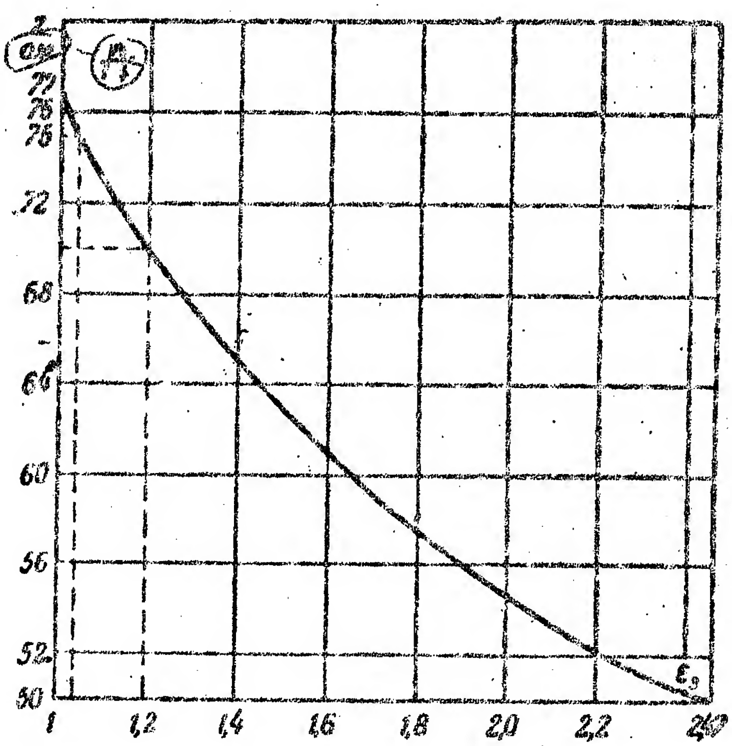


Fig. 5-17. The Wave Impedance of a Cable as a Function of the Permitivity of the Insulation where D/d = 3.6. A) Ohms.

An increase in D/d leads to a logarithmic rise in the wave impedance.

b. Phase Constant, Rate of Propagation. With the aid

of the same formulas (5-12) and (5-13), the phase constant of a coaxial cable may be determined:

$$a = \omega\sqrt{LC} = \frac{\omega\sqrt{\varepsilon}}{300\,000} = \frac{\omega\sqrt{\varepsilon}}{c} \; [\text{rad}\ /\text{км}], \tag{5-16}$$

where c = 300,000 km/sec is the velocity of propagation of energy in the air.

In turn, the velocity of propagation of electromagnetic energy along coaxial cables may be computed from the formula:

$$v = \frac{\omega}{a} = \frac{300\,000}{\sqrt{\varepsilon}} = \frac{c}{\sqrt{\varepsilon}}. \tag{5-17}$$

As the frequency rises, the phase constant increases linearly. This is due to the fact that the rate at which energy is transmitted along a coaxial cable is almost completely constant over the entire frequency band under consideration. The velocity increases as the dielectric constant ε becomes larger. Thus, in a cable with a solid dielectric (ε = 2.3), v = 200,000 km/sec, while in a cable with combination insulation (ε = 1.1), v = 280,000 km/sec.

Energy is transmitted at a higher rate over coaxial cables than over other types of cables, and the propagational velocity approaches the velocity of the electro-

magnetic waves in air, i.e., c ≈ 300,000 km/sec.

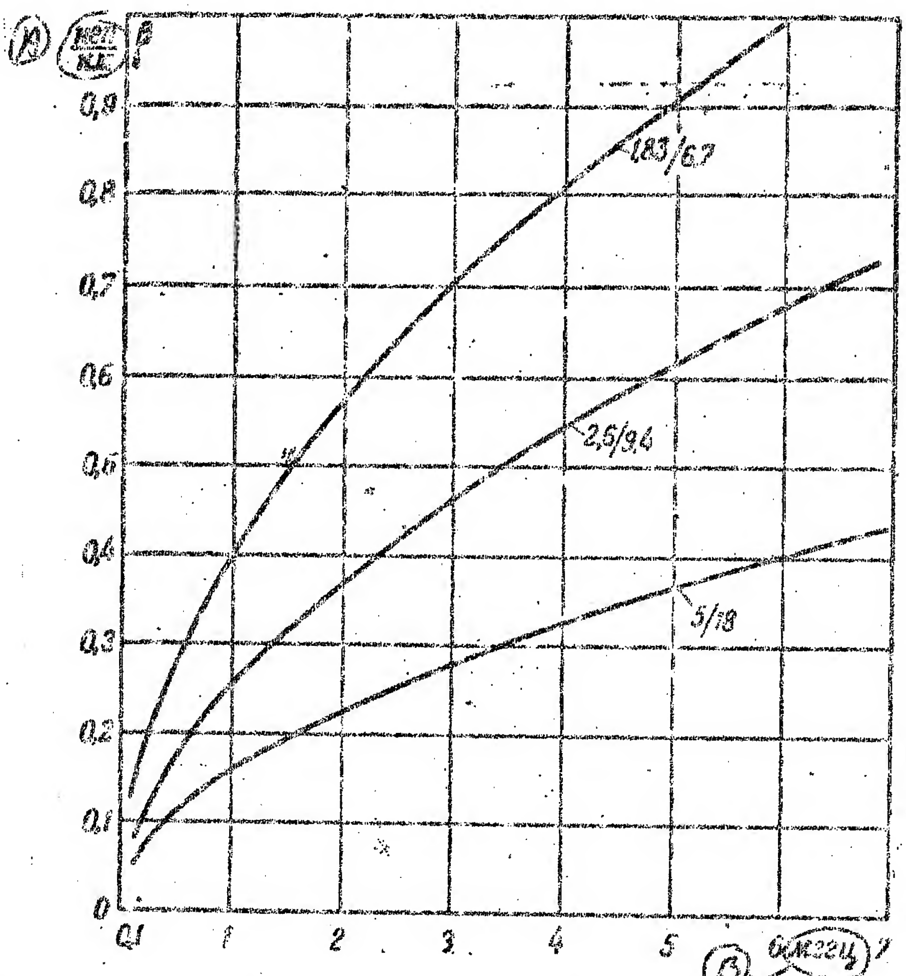


Fig. 5-18. The Attenuation of Various Types of Cables as a Function of Frequency. A) Nepers/km; B) Mc.

Type of Cable	Z, ohms	C, mμf/km
1.83/6.7	75	47
2.6/9.4	74	48
5/18	70	50

c. Wave Length Decrease Factor. In high-frequency communications work, the electro-magnetic wave length decrease factor ξ is widely used to evaluate propagation.

This factor characterizes the degree to which the propagation velocity of electro-magnetic energy is decreased in the cable in comparison to its velocity in air

$$\xi = \frac{c}{v} = \frac{300\,000\sqrt{\varepsilon}}{300\,000} = \sqrt{\varepsilon}. \qquad (5\text{-}18)$$

It follows from formula (5-18) that as the dielectric constant of the cable increases, the coefficient also increases.

In a cable having solid polyethylene insulation ($\varepsilon = 2.3$), $\xi = 1.52$, and consequently the velocity of propagation of energy along such a line is 1.52 times less than in air.

d. Cable attenuation. It follows from Table 5-5 and Fig. 5-16 that as the frequency increases, the cable attenuation β rises; in modern cables having a high-quality dielectric, the attenuation increases with the $\sqrt{f}$.

Fig. 5-18 gives the frequency dependence of attenuation for cables of the most common types (5/18; 2.6/9.4, and 1.83/6.7); it follows from the figure that the attenuation is inversely proportional to the dimensions of the cable.

Thus, the attenuation of a 2.6/9.4 cable is greater

than the attenuation of a 5/18 cable by 18/9.4 = 1.9 times.

Below we consider β as a function of the ratio D/d.

Tables 5-6 and 5-7 give the electrical characteristics of large (5/18) and small (1.83/6.7) trunk coaxial cables.

A f, гц	B R, ом/км	C L, мгн/км	D C, нф/км	E G, мкмо/км	F β, мнеп/км	G α, рад/км	H z, ом	I v, км/сек
$6\cdot10^4$	14,1	0,247	48	16	89,6	1,42	79	$2,63\cdot10^5$
10^5	18,5	0,287	48	26,3	130	2,32	77	$2,68\cdot10^5$
10^6	58	0,268	48	253	400	22,2	74,2	$2,83\cdot10^5$
$3\cdot10^6$	99	0,264	48	800	704	66	74	$2,85\cdot10^5$
$7\cdot10^6$	153	0,263	48	1 865	1 120	153	73,7	$2,89\cdot10^5$
10^7	184	0,262	48	2 630	1 345	217,6	73,6	$2,9\cdot10^5$

TABLE 5-6. Electrical Characteristics of a Small Coaxial Cable (1.83/6.7). A) f, cps; B) R, ohms/km; C) L, mh/km; D) c, $m\mu$f/km; E) G, μmho/km; F) β, m neper/km; G) α, rad/km; H) z, ohm; I) v, km/sec.

(A) f, гц	(B) R, ом/км	(C) L, мгн/км	(D) C, нф/км	(E) G, мкмо/км	(F) β, мкнеп/км	(G) α, рад/км	(H) z, ом	(I) v км/сек
$6\cdot10^{4}$	5,12	0,273	53	1,35	36	1,41	71,5	$2{,}63\cdot10^{5}$
10^{5}	6,72	0,267	53	2,25	47,5	2,36	71,0	$2{,}65\cdot10^{5}$
10^{6}	21,3	0,259	53	22,5	153	23	70	$2{,}73\cdot15^{6}$
$3\cdot10^{6}$	36,5	0,257	53	67,5	265	68,5	69,6	$2{,}75\cdot10^{5}$
$7\cdot10^{6}$	56,5	0,257	53	157,5	410	159	69,1	$2{,}79\cdot10^{5}$
10^{7}	67,5	0,255	53	225	490	225	69,0	$2{,}8\cdot10^{5}$

TABLE 5-7. Electrical Characteristics of a Large Coaxial Cable (5/18). A) f, cps; B) R, ohms/km; C) L, mh/km; D) c, mμf/km; E) G, μmho/km; F) β, m neper/km; G) α, rad/km; H) z, ohm; I) v, km/sec.

5-6. Calculating the Resistance of an Outer Conductor of Complex Design.

It was shown above that in order to achieve the required flexibility in a coaxial cable, and also on the basis of production considerations the outer conductor is made not as a continuous tube, but rather as a fairly complex structure.

The structural form of the conductor has a considerable effect upon its resistance, and consequently, on

the coaxial cable attenuation.

We will consider the design of a coaxial cable having an outer conductor formed of spirally wound strips, and of tubes with stamped couplings.

I. SPIRAL OUTER CONDUCTOR

In this type of a structure (Fig. 5-19) the current in the outer conductor does not flow along the axis of the cable, but rather follows a line of a screw-thread, whose angle of advance corresponds to the angle of advance of the axis of the tape.

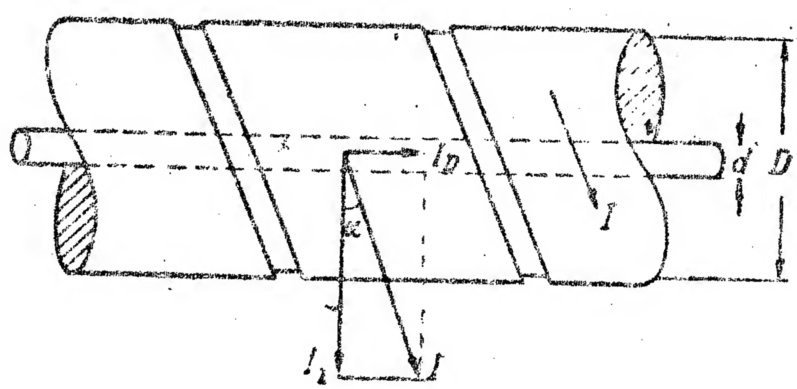


Fig. 5-19. Computing the Resistance of a Spiral Outer Conductor.

As may be seen from the figure, the current in the

outer conductor may be considered to have two components.

I_D--the current along the axis of the cable;

I_z--the current in the lateral direction.

The current I_z causes a magnetic field H_z having an axial (longitudinal) direction. The field induces eddy currents along the outer conductor and other shells of the cable, causing additional losses, and leading to an increase in the resistance of the outer conductor of the coaxial cable by a magnitude R_z. The formula for computing R_z, which has been verified both theoretically and experimentally, takes the following form:

$$R_z = \sqrt{\frac{\mu f}{7}} \cdot \frac{2}{\sqrt{10}} \cdot \frac{1}{D} \cot^2 \alpha \, [\text{ohm}/\kappa\mathcal{M}], \qquad (5\text{-}19)$$

When copper tape is utilized ($\mu = 1$) the formula is simplified:

$$R_z = 0{,}0835\sqrt{f}\frac{1}{D}\cot^2 \alpha \, [\text{ohm}/\kappa\mathcal{M}], \qquad (5\text{-}20)$$

where D is the inside diameter of the outer conductor, mm.

Comparing formulas (5-9) and (5-20), we notice that R_D and R_z differ only in the coefficient $\cot^2 \alpha$, which takes into account the effect of the spiral structure of the outer conductor

$$R_z = R_D \cot^2 \alpha, \tag{5-21}$$

The additional resistance introduced into the cable circuit is determined by the law of variation of $\cot^2 \alpha$, and depends on the angle of advance of the spiral.

The larger this angle, the less the additional longitudinal magnetic field H_z, and the lower the transmission losses. In the limiting case $\alpha = 90^\circ$, the spiral is equivalent to a longitudinal strip and the additional resistance is $R_z = 0$.

At $\alpha = 45^\circ$, the resistances R_d and R_z are equal, since in this case the current is distributed equally between the longitudinal and lateral directions ($I_D = I_z$).

The total resistance of a cable having a spiral-type outer conductor is expressed by the equation:

$$R = R_d + R_D + R_z = 0{,}0835\sqrt{f}\left(\frac{1}{d} + \frac{1}{D} + \frac{1}{D}\cot^2 \alpha\right) [\text{ohm}/\kappa\text{м}]. \tag{5-22}$$

Fig. 5-20 gives you the calculated values of the resistance R_z at $f = 10^6$ cps for a type 5/18 cable having a spiral outer conductor, for different angles α.

For the sake of comparison the values of R_d and R_D are also given.

Table 5-8 illustrates the frequency dependence of the cable resistance for a spiral-type outer conductor.

If there is any sort of metallic shell above a spiral-type outer conductor, owing to the longitudinal magnetic field eddy currents will be introduced into the shell as well; this causes additional losses: in this case the resistance is computed according to the formula:

$$R_s = \left(R_D + \sqrt{\frac{\pi f}{\gamma}} \cdot \frac{2}{\sqrt{10}D} \right) \cot^2 \alpha \ [\text{ohm/km}], \tag{5-23}$$

where $R_D \cot^2 \alpha$ allows for the losses in the spiral-type outer conductor, and the second term takes care of the losses in the surrounding metallic shell.

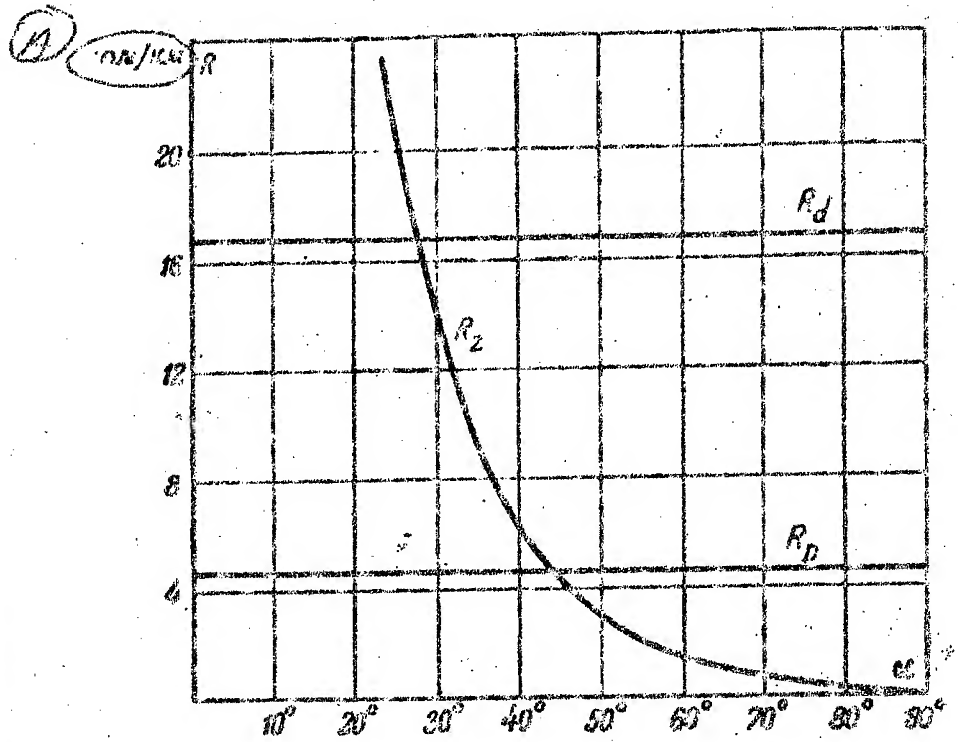


Fig. 5-20. The Dependence of the Resistance of a Spiral Outer Conductor, R_z, on the Angle of Advance of the Spiral, α. A) Ohm/km.

TABLE 5-8.

Внешний провод кабеля (A)	Наименование параметра (B)	Частота, гц (C)					
		10^5	10^6	10^7	10^8	10^9	$3\cdot10^9$
Идеальная конструкция (D)	R_d	5,28	16,8	52,8	168,0	528	920
	R_D	1,47	4,65	14,7	46,5	147	254
	$R=R_d+R_D$	6,75	21,45	67,5	214,5	675	1 174
(E) Спиральная лента $\alpha=25^\circ$	R_z	6,8	21,5	68	215	680	1 180
	$R=R_d+R_D+R_z$	13,55	42,95	135,5	429,5	1 355	2 354
(F) Спиральная лента $\alpha=30^\circ$	R_z	4,44	14	44,4	140	444	770
	R	11,19	35,45	111,9	354,5	1 119	1 944
(G) Спиральная лента $\alpha=45^\circ$	R_z	1,47	4,65	14,7	46,5	147	254
	R	8,22	26,1	82,3	261	822	1 428
(H) Спиральная лента $\alpha=60^\circ$	R_z	0,5	1,56	5	15,6	50	85,6
	R	7,25	23,01	72,5	230,1	72,5	1 259,6
(I) Спиральная лента $\alpha=80^\circ$	R_z	0,046	0,145	0,46	1,45	4,6	8,0
	R	6,8	21,6	67,96	216	679,6	1 182

The Frequency Dependence of the Resistance of a Type 5/18 Cable having a Spiral-Type Outer Conductor.

A) Outer Conductor of the Cable; B) Parameter; C) Frequency, cps; D) Ideal Structure; E) Spiral Tape, $\alpha = 25^\circ$; F) Spiral Tape, $\alpha = 30^\circ$; G) Spiral Tape, $\alpha = 45^\circ$; H) Spiral Tape, $\alpha = 60^\circ$; I) Spiral tape, $\alpha = 80^\circ$.

The values of μ and γ' correspond to the material of the surrounding shell: Lead, etc.

It should be kept in mind, that owing to the additional losses introduced by the spiral nature of the outer conductor, the optimum relationship of the diameters of the inner and outer conductors which was given above (3.6) will change somewhat.

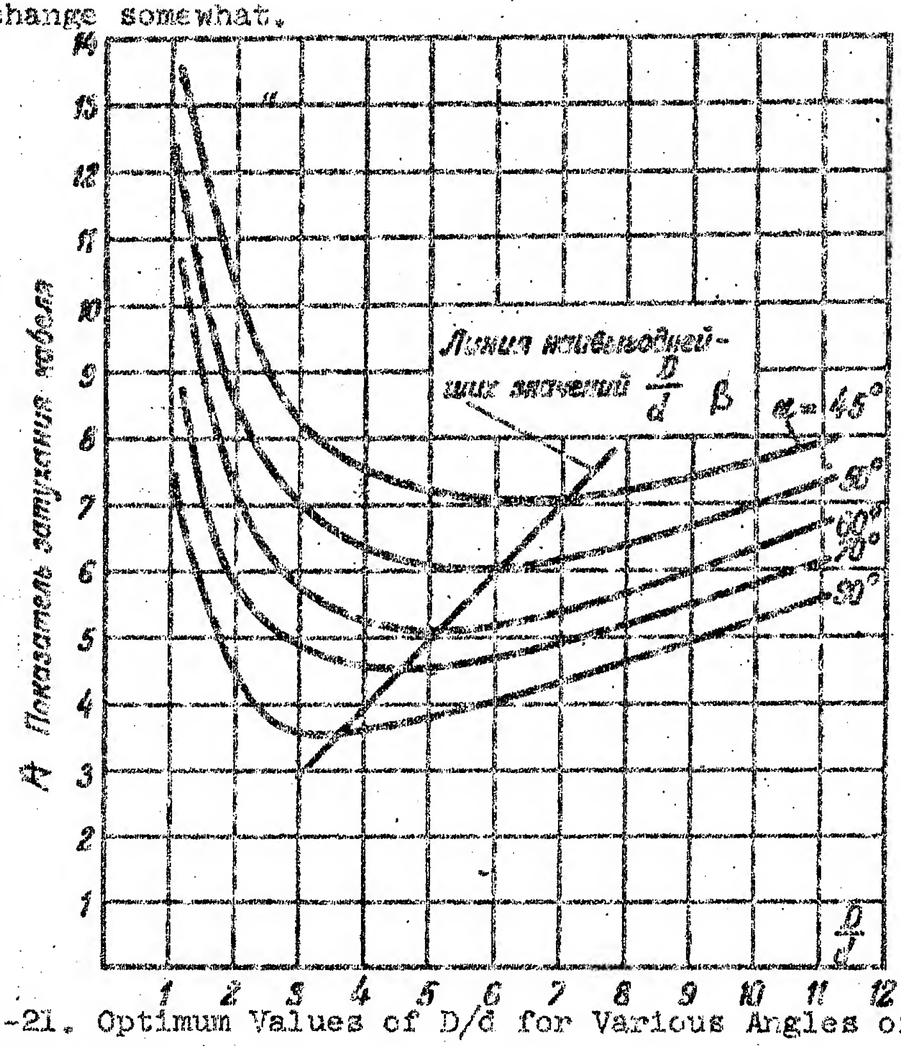


Fig. 5-21. Optimum Values of D/d for Various Angles of Advance of a Spiral. A) Degree of cable attenuation; B) Line of most favorable values of D/d.

Fig. 5-21 gives the optimum values for the diameters of the outer, D, and inner, d, copper conductors, for various angles of advance of the tape, α, and a surrounding lead sheath. It is clear from the figure, that as the amount of spiral in the conductor decreases from 90° to 45°, the ratio D/d rises from 3.6 ($\alpha = 90^{\circ}$ and there is no longitudinal magnetic field) to 7.

If, in a cable with copper conductors, where D/d 3.6, the cylindrical (ideal) outer conductor is replaced with a spiral-type conductor, the increase in attenuation may be expressed, using the spiral angle, by the following simple formula:

$$\beta = \beta_0 \frac{1}{\sin^2 \alpha} = \frac{2.5\sqrt{f_c} \cdot 10^{-4}}{D} \cdot \frac{1}{\sin^2 \alpha} \text{ [nepers/km]} \quad (5\text{-}24)$$

where β_0 is the optimum attenuation of the cable with an ideal outer conductor [see formula (5-34)].

It follows from expression (5-24) that the additional attenuation introduced by the spiral follows a $1/\sin^2\alpha$ law.

For large values of α (the angle is measured from a line perpendicular to the axis of the cable), the losses owing to the spiral are comparatively small, while as α decreases, the losses rise sharply.

Thus, at $\alpha = 70^\circ$, the attenuation is increased by 13%; at $\alpha = 60^\circ$ by 34%; while at angles of less than $\alpha = 45^\circ$, the attenuation is doubled (in comparison with the attenuation of an ideal conductor). Thus, in designing a cable with a spiral-type outer conductor, it is necessary to make the spiral as long as possible.

The pitch of the spiral, however, is also affected by considerations of mechanical strength, cable flexibility, manufacturing considerations, and the final choice will represent a compromise among all of these requirements.

II. OUTER CONDUCTOR FORMED BY STAMPED HALF-TUBES WITH GROOVES

Just as in the previous case, this design (Fig. 5-22) violates the basic principle of an ideal outer sheath--the absence of an external magnetic field.

The lateral grooves distort the magnetic field, and a longitudinal component appears to increase the energy losses; this is observed as an additional resistance R_z.

The value of R_z is computed with the aid of the empirical formula given below; it is in complete agreement with the results of measurements of similar types of cables:

$$R_z = \sqrt{\frac{\mu f}{\gamma}} \cdot \frac{2}{\sqrt{10}} \cdot \frac{1}{D'} \cdot \frac{a+b}{l} \, [\text{ohm}/\text{км}], \tag{5-25}$$

where a is the width of a channel;

b is the depth of a channel;

l is the distance between channels;

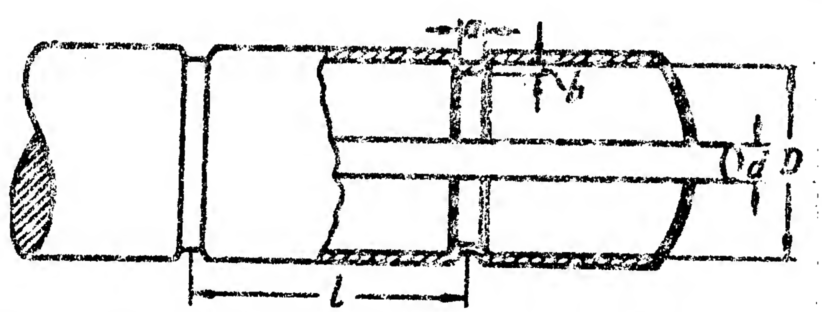
$D' = D-2b$ (all dimensions in mm).

Or, if the outer conductor is made of copper,

$$R_z = 0{,}0635 \sqrt{f} \, \frac{1}{D'} \cdot \frac{a+b}{l} \, [\text{ohm}/\text{км}]. \tag{5-26}$$

It is clear from the formula that the additional resistance increases with the width of the channel. The resistance decreases where the channels are located infrequently along the cable. In order to provide the required cable flexibility, channels must be located every 20-30 mm.

Fig. 5-22.



Calculating the Resistance of an Outer Conductor Made of Stamped Half-Tubes with Grooves.

The complete formula for calculating a coaxial cable having an outer conductor of copper half-tubes with grooves takes the following form:

$$R = R_d + R_D + R_z = 0{,}0835\left(\frac{1}{d} + \frac{1}{D} + \frac{1}{D'}\cdot\frac{a+b}{t}\right)\text{ohm}/\kappa M], \tag{5-27}$$

It may also be used to calculate the resistance of a tubular outer conductor.

Table 5-9 gives the results of calculating the resistance of type 5/18 cable with an outer conductor of stamped copper half-tubes with grooves.

TABLE 5-9.

Resistance of 5/18 Cable with an Outer Conductor of Stamped Copper Half-Tubes with Grooves (in ohm/km).

(B) f, гц / (A) Тип внешнего провода кабеля	10^5	10^6	10^7	10^8	10^9	$3\cdot 10^9$
Полутрубки с перехватами (C) . .	6,95	22	69,5	220	695	1 210
Идеальная конструкция (D) .	6,75	21,45	67,5	214,5	675	1 174
Спиральная лента ($\alpha = 45°$) (E) . .	8,22	26,1	82,3	261	822	1 423

A) Type of outer conductor; B) f, cps; C) Half-tubes with grooves; D) Ideal design; E) Spiral tape ($\alpha = 45°$).

In the calculation, it is assumed that $a = 2$ mm, $b = 1.5$ mm, $l = 28$ mm.

For the sake of comparison the values of R are also given for cables with spiral-tape type outer conductors ($\alpha = 45°$) and the ideal design.

It follows from Table 5-9 that the resistance of an outer conductor of stamped half-tubes with grooves is very little (up to 3%) above the resistance of a cable conductor of ideal design. When a spiral-type outer conductor is used, the resistance increases by 21%.

5-7. The Minimum Attenuation of a Coaxial Cable and the Communication Span

Since it is desirable that the relationship

$$v = \frac{1}{\sqrt{LC}} = \lambda f$$

(λ is the wave length) hold for coaxial cables used in the high-frequency band, the expression for the attenuation reduces to the following form:

$$\beta = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} = \frac{1}{2\lambda f}\left(\frac{R}{L} + \frac{G}{C}\right) = \frac{\pi}{\lambda}\left(\frac{R}{\omega L} + \frac{G}{\omega C}\right). \tag{5-28}$$

Here the quantities $R/\omega L = \tan \varepsilon$ and $G/\omega C = \tan \delta$ are the metal and dielectric loss factors. They are determined by measurements of short sections of cable which are short-circuited or insulated at the end.

Thus, the attenuation may be represented as the sum of the attenuations owing to losses in metal and in the dielectric:

$$\beta = \beta_R + \beta_G = \frac{\pi}{\lambda}\tan \varepsilon + \frac{\pi}{\lambda}\tan \delta, \tag{5-29}$$

where the ratio of β_G to β_R = $\tan \delta : \tan \varepsilon$.

In trunk coaxial cables with high-frequency insulation, the losses in the dielectric are considerably

less than the losses in the metal. In particular, in modern cables tan ε is roughly 10^{-2}, while tan δ is only 10^{-4}. Accordingly, the attenuation owing to losses in the dielectric do not exceed several per cent of the total cable attenuation.

In order to find the nature of the variation in the attenuation of a coaxial cable as a function of its dimensional relationships (d and D) and the quality of the basic materials (tan δ and ε), we substitute the value of the primary parameters into the expression for the attenuation. After some manipulations we obtain:

$$\beta = \beta_R + \beta_G = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} = \frac{8{,}35\sqrt{f_k}\left(\frac{D}{d}+1\right)\cdot 10^{-3}}{12D\ln\frac{D}{d}} + \frac{10}{3}\pi f\sqrt{\varepsilon}\tan\delta\cdot 10^{-6}. \quad (5\text{-}30)$$

It has been shown before that in modern coaxial cables the losses in the dielectric are negligible, and thus it is possible to assume that the attenuation of a cable may be determined accurately enough by means of the losses in the metal:

$$\beta = \beta_R = \frac{R}{2}\sqrt{\frac{C}{L}} = \frac{8{,}35\sqrt{f_k}\left(\frac{D}{d}+1\right)\cdot 10^{-3}}{12D\ln\frac{D}{d}}. \quad (5\text{-}31)$$

It follows from expression (5-31) that as D/d increases, its numerator increases, while the denominator decreases logarithmically. This gives reason to suppose that at a specific ratio of cable dimensions, the attenuation will reach a minimum. The existence of such an optimum ratio follows directly from the formula:

$$\beta = \frac{R}{2}\sqrt{\frac{C}{L}},$$

where each of the parameters R, L, and C depend upon D/d. Investigating the function for its minimum as a function of D/d, we find that β is minimal for D/d = 3.6, which should be kept in mind when designing coaxial cables.

This is correct for cables with copper conductors. Where they are manufactured from different metals, the condition for minimum attenuation will fall at another D/d ratio.

In this case the minimum attenuation may be determined by means of the following expression:

$$\ln\frac{D}{d} = 1 + \frac{R_D}{R_d}, \tag{5-32}$$

where R_D and R_d are, respectively, the resistance of the outer and inner conductors.

The relationship R_D/R_d may be written in the form

$$\frac{R_D}{R_d} = \sqrt{\frac{\mu_D \gamma_d}{\mu_d \gamma_D}} \cdot \frac{d}{D},$$

or, since for the materials used in manufacturing cables, $\mu = 1$:

$$\frac{R_D}{R_d} = \sqrt{\frac{\gamma_d}{\gamma_D}} \cdot \frac{d}{D}.$$

The expression (5-32) then takes the form:

$$\ln \frac{D}{d} = 1 + \frac{d}{D} \sqrt{\frac{\gamma_d}{\gamma_D}}, \tag{5-33}$$

where γ_d and γ_D are, respectively, the conductances of the metals of the inner and outer conductors.

Where γ_D and γ_d are equal, the optimum condition will be: $\ln \frac{D}{d} = 1 + \frac{d}{D}$.

In this case the minimum attenuation will occur at $D/d = 3.6$. Below we have given the optimum ratio D/d for various types of cables, assuming in all cases that the inner conductor is made of copper, while the outer conductor is made of the material shown in the table.

TABLE

	Copper	Aluminum	Iron	Lead	Zinc
$\frac{D}{d}$	3,6	3,9	4,2	5,2	4,3

From the graph of Fig. 5-23, β as a function of D/d for various types of cable, it is clear that if the optimum relationship is disturbed (especially if the ratio is too low) there will be a rather sharp increase in the attenuation.

Keeping in mind the necessity of adhering to the most favorable dimensional relationship, we substitute these dimensions into formula (5-31), and obtain the formula for the attenuation of a coaxial cable having copper conductors:

$$\beta = \beta_R = \frac{2{,}5\sqrt{f\varepsilon} \cdot 10^{-6}}{D} \text{ [nepers/km]} \qquad (5\text{-}34)$$

from which it follows that the attenuation increases as f and ε rise and drops sharply as the diameter of the cable goes up.

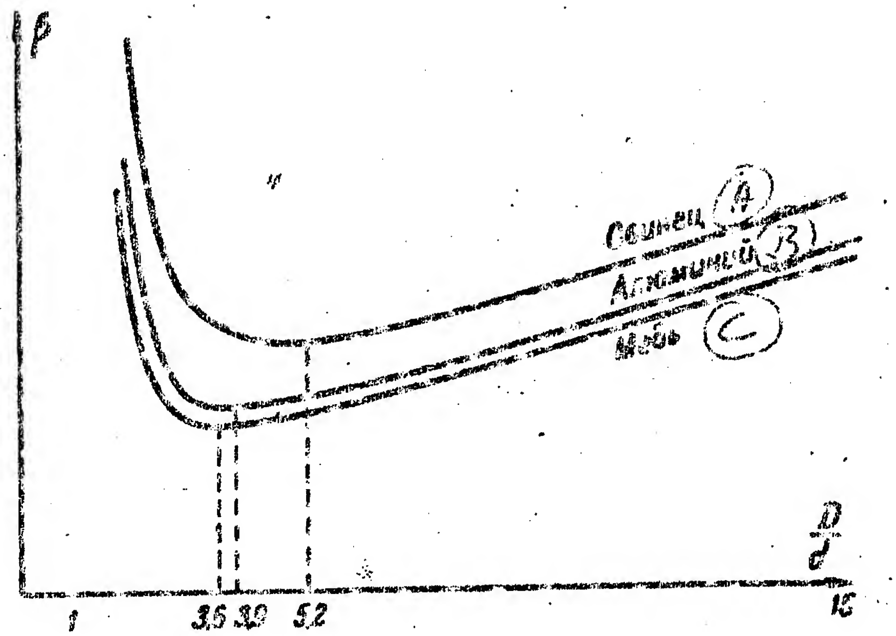


Fig. 5-23. Curve of the Attenuation of Cables of Different Metals As a Function of the Variation in the Ratio D/d. A) Lead; B) Aluminum; C) Copper.

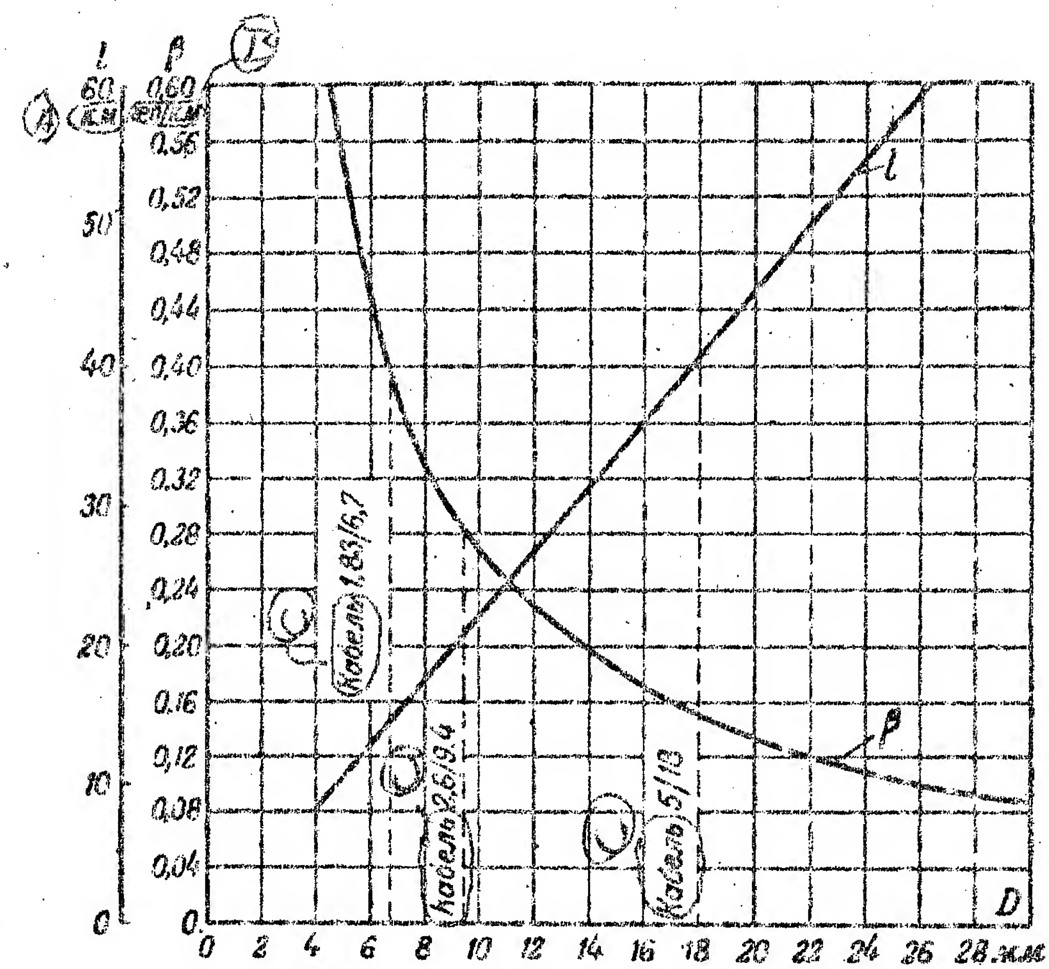


Fig. 5-24. Variation in Cable Attenuation and Communication Span with Increasing Outside Diameter of the Cable ($f = 10^6$ cps, $b = 6$ nepers). A) km; B) nepers/km; C) cable.

Figure 5-24 shows the attenuation of the most common types of trunk cables and the general nature of their variation with increasing D. The calculation was carried out at $f = 10^6$ cps.

The wave impedance for a cable with the optimum ratio of diameters ($D/d = 3.6$) is expressed by the following formula:

$$z = 77/\sqrt{\varepsilon}\,[\text{ohm}]. \qquad (5\text{-}35)$$

A tentative choice of cable for a trunk link can be made with the aid of formula (5-34).

Keeping in mind that modern repeater equipment compensates for an attenuation of about 6-7 nepers, the cable attenuation constant β may be related to the distance between amplifier points, l, by the following equation:

$$\beta = \frac{b}{l} = \frac{6-7}{l}\,[\text{nepers/km}]$$

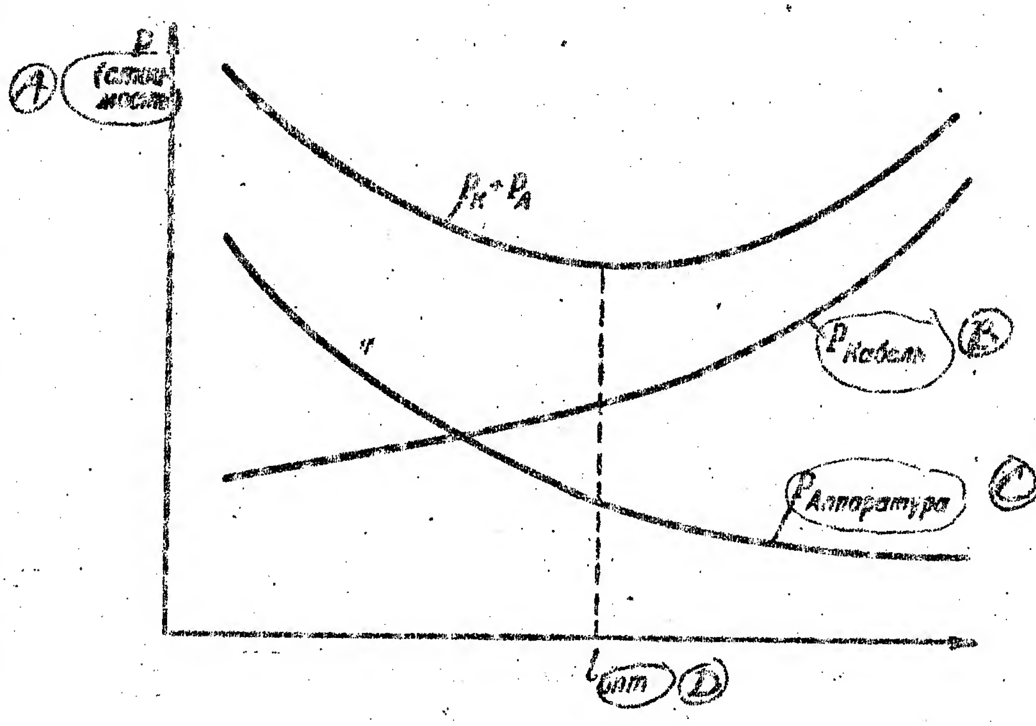


Fig. 5-25. Cost of Cable and Equipment for Various Distances Between Repeater Points. A) Cost; B) P_{cable}; C) $P_{equipment}$; D) opt.

Expression (5-34) may be given in the following form

$$D = \frac{2{,}5\sqrt{f}\epsilon l}{b} \cdot 10^{-4}, \qquad (5\text{-}36)$$

where b is the gain of the equipment, or (what is the same

thing in this case) the cable attenuation, equal to 6-7 nepers.

Consequently, the inside diameter of the outer conductor of the cable, D, is determined from the given communication span l (the distance between repeater points), and the frequency range, and from the value of D, the optimum value of the inner diameter of the cable is computed using the relationship $d = D/3.6$.

The calculation is carried out using the upper limit of the transmitted frequency band.

In this manner the geometric dimensions and type of cable are determined which will guarantee the passage of the required frequency band for the required distance.

Fig. 5-24 shows the dimensions of a coaxial cable required to cover the corresponding distance between amplifier points.

In turn, the most favorable repeater-point separation, l, is chosen on the basis of technical and economic analyses of the costs of equipment and cable.

Fig. 5-25 gives the generalized dependences of the cost of equipment and cable for various repeater-point separations; it follows from the figure that as l increases, the expenditure on equipment will decrease, since fewer repeaters will be required along the trunk. The cost of

the cable will rise, since it will be necessary to increase its dimensions with a greater direct-link span. On the graph, the optimum distance between repeater points corresponds to the minimum total expenditure on the cable trunk.

5-8. Determining the Equivalent Values of ε and tan δ for a Coaxial Cable.

In designing a cable it is necessary to know the resultant so-called equivalent values of the dielectric constant ε_e and the dielectric loss angle tan δ_e .

The determination is complicated by the fact that the structure of the insulating layer of modern coaxial cables is fairly complicated. The insulation takes various forms (supporting spirals, caps, etc.) and may consist of differing dielectrics.

For a cable with solid insulation, the dielectric constant ε_e and the dielectric loss angle δ_e will equal, respectively, the ε and tan δ of the materials from which the insulation is made, i.e.,

$$\varepsilon_э = \varepsilon \ , \ \tan\delta_э = \tan\delta.$$

For combination insulation, the calculation of ε_e and tan δ_e is considerably complicated.

Using the theory of a cylindrical capacitor with multilayer insulation, it may be shown that the resultant

values of the dielectric constant and the dielectric loss angle for combination insulation are determined by the following expressions:

$$\varepsilon_e = \frac{\varepsilon_1 V_1 + \varepsilon_2 V_2}{V_1 + V_2}, \tag{5-37}$$

$$\tan \delta = \frac{\varepsilon_1 V_1 \tan \delta_1 + \varepsilon_2 V_2 \tan \delta_2}{\varepsilon_1 V_1 + \varepsilon_2 V_2}, \tag{5-38}$$

where the values with the index 1 correspond to the first dielectric, and the values for the index 2 to the second dielectric.

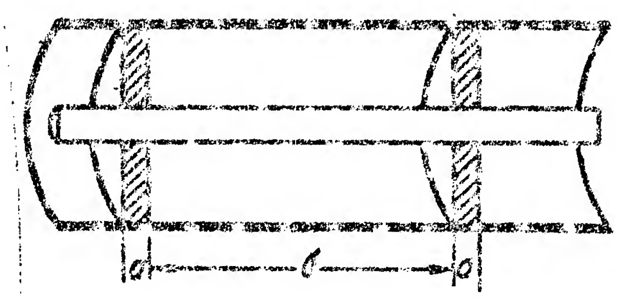


Fig. 5-26. Calculating ε_e and $\tan \delta_e$ For a Bead-Insulated Cable.

Since, structurally speaking, the insulation takes the form of a uniform elementary section periodically repeated along the length of the cable, it is possible to carry out a calculation of ε_e and $\tan \delta_e$

for a unit length of the cable, and to substitute the areas q for the volumes V in the formulas given above.

Where the insulation is a combination of air (ε_a and $\tan\delta_a$) and dielectric (ε_d and $\tan\delta_d$), the formulas may be rewritten as follows:

$$\varepsilon_e = \frac{\varepsilon_a V_a + \varepsilon_d V_d}{V_a + V_d}; \tag{5-39}$$

$$\tan\delta_e = \frac{\varepsilon_d V_d}{\varepsilon_a V_a + \varepsilon_d V_d} \tan\delta_d. \tag{5-40}$$

Using these formulas, it is not difficult to obtain the values of the resultants ε_e and $\tan\delta_e$ for discontinuous bead-type insulation (Fig. 5-26):

$$\varepsilon_e = \frac{\varepsilon_a b + \varepsilon_d a}{a + b} \tag{5-41}$$

$$\tan\delta_e = \frac{\varepsilon_d a}{\varepsilon_a b + \varepsilon_d a}, \tag{5-42}$$

where b is the distance between beads;

a is the thickness of a bead.

Calculation of the electrical characteristics of combination insulation of continuous type involves serious

difficulties. The reason for this is that it is necessary to determine not only the relationship of the volumes of dielectric and air, but also to evaluate the effect of their relative location on the resultant properties ε_e and $\tan \delta_e$.

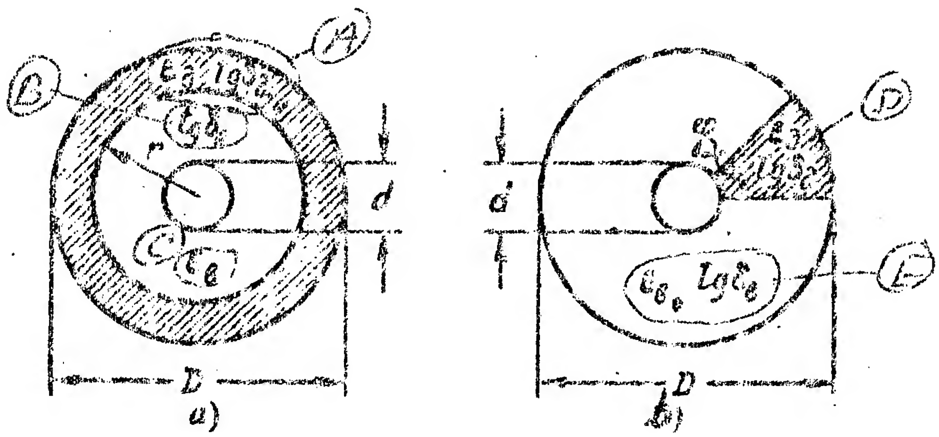


Fig. 5-27. Calculating Two-Layer Insulation For a Coaxial Cable. a) Distribution of the layers in the radial direction; b) Distribution of the layers in the tangential direction. A) ε_d $\tan \delta_d$; B) $\tan \delta_a$; C) ε_a; D) ε_d $\tan \delta_d$; E) ε_a $\tan \delta_a$.

It is very desirable that the method of calculation utilize an arbitrary substitution of an equivalent insulation for the actual insulation of the cable so as to reduce the problem to one of the two-layer forms of insu-

lation, which have been analyzed mathematically. Here we use the designations shown in Fig. 5-27 for combinations of layers in the radial (r) direction and in the tangential (φ) direction.

For radially combined insulation, the equivalent values ε_e and $\tan\delta_e$ will be:

$$\varepsilon''_{er} = \frac{\varepsilon_1\varepsilon_2 \ln\frac{D}{d}}{\varepsilon_2 \ln\frac{d_r}{d} + \varepsilon_1 \ln\frac{D}{d_r}}, \tag{5-43}$$

$$\tan\delta_{er} = \frac{\tan\delta_1\varepsilon_2 \ln\frac{d_r}{d} + \tan\delta_2\varepsilon_1 \ln\frac{D}{d_r}}{\varepsilon_2 \ln\frac{d_r}{d} + \varepsilon_1 \ln\frac{D}{d_r}}, \tag{5-44}$$

where d_r is the diameter of the boundary between the various dielectric media.

The quantities with the index 1 refer to the first dielectric, those with the index 2 – to the second.

Combination insulation most often consists of air (ε_a and $\tan\delta_a$) with a dielectric (ε_d and $\tan\delta_d$). In this case, formulas (5-43) and (5-44) take the form:

$$\varepsilon_{er} = \frac{\varepsilon_a \varepsilon_d \ln \frac{D}{d}}{\varepsilon_d \ln \frac{d_r}{d} + \varepsilon_a \ln \frac{D}{d_r}}, \tag{5-45}$$

$$\tan \delta_{er} = \frac{\varepsilon_a \ln \frac{D}{d_r}}{\varepsilon_d \ln \frac{d_r}{d} + \varepsilon_a \ln \frac{D}{d_r}} \tan \delta_d. \tag{5-46}$$

For two-layer tangential combination insulation of air and dielectric, the equivalent values ε_e and $\tan \delta_e$ are determined from the following expressions:

$$\varepsilon_{e\varphi} = \varepsilon_a + (\varepsilon_d - \varepsilon_a) \frac{\varphi}{2\pi}, \tag{5-47}$$

$$\tan \delta_{e\varphi} = \frac{\varepsilon_d \varphi}{2\pi\varepsilon_a + (\varepsilon_d - \varepsilon_a)\varphi} \tan \delta_d, \tag{5-48}$$

where φ is an angle that characterizes the proportion of dielectric in the total cross section of the cable.

In order to compute the equivalent values ε_e and $\tan \delta_e$ from the formulas given above, it is necessary to know the proportion of the dielectric in the total cross section of the cable; this is determined by means of the parameters d_r and φ. This cannot always be established with an adequate degree of accuracy: the value ε_e may be obtained

from the results of measurement of the cable capacitance, but the parameter tan δ_e is very difficult to measure (especially at high frequencies), since its absolute value is exceptionally small.

For this reason, the following method is used in practice to determine the equivalent values of the parameters ε_e and tan δ_e.

The parameter ε_e is determined on the basis of the result of a measurement of the cable capacitance; then a comparison is made of the dielectric constant of two specimens of cable, one with the combination insulation, and the other with solid insulation: then r or φ, and consequently, the relation of the volumes (areas) of the dielectric and air in the cable are found.

On the basis of the values of r and φ which are found, using formulas (5-46) and (5-48), tan δ_{er} and tan $\delta_{e\varphi}$ are found.

If we substitute the values of r and φ, expressed as functions of the dielectric constants, into formulas (5-46) and (5-48), we will obtain

$$\tan\delta_{er} = \frac{\varepsilon_e - \varepsilon_a}{\varepsilon_d - \varepsilon_a}\tan\delta_d, \tag{5-49}$$

$$\tan\delta_{e\varphi} = \frac{\varepsilon_e - \varepsilon_a}{\varepsilon_d - \varepsilon_a}\tan\delta_d\,\frac{\varepsilon_d}{\varepsilon_e}. \tag{5-50}$$

The calculation of $\tan \varphi$ is carried out using the formula corresponding to the arbitrarily chosen type of insulation (radial or tangential), to which the form of the actual insulation approximates.

If the structure of the insulation is so complicated that it is not possible to determine the type of arbitrary substitution to be made, then in order to determine the dielectric loss angle, the radial and tangential magnitudes are averaged:

$$\tan \delta_e = \frac{\tan \delta_{er} + \tan \delta_{e\varphi}}{2} = \frac{\varepsilon_d + \varepsilon_a}{2\varepsilon_e} \cdot \frac{\varepsilon_e - \varepsilon_a}{\varepsilon_d - \varepsilon_a} \tan \delta_d. \quad (5\text{-}51)$$

Formula (5-51) is very widely used in practice, and gives good agreement with experimental data.

We compare the values of dielectric loss angles for radial, $\tan \delta_r$, and tangential, $\tan \delta_\varphi$, combination-insulation constructions; for the tangential dielectric arrangement the losses will be greater by $\tan \delta_\varphi / \tan \delta_r = \varepsilon_d / \varepsilon_e$ times. For various types of dielectrics, this ratio equals 2-4.

This confirms the fact that in order to calculate ε_e and $\tan \delta_e$ for complex dielectrics, it is necessary to know not only the ratio of the volumes of the components (air-dielectric), but also to allow for the disposition of

the insulation over the cross section of the cable. In principle, this is associated with the direction of the electrical field of a coaxial cable.

As is known, the lines of force of the electrical field of a coaxial cable are radial in direction. Thus the greatest concentration of lines of force occurs at the center of the system — about the inner conductor of the coaxial cable.

The further from the center of the cable, the less dense the lines of force of the electrical field.

It is clear, that if the greatest concentration of the field occurs within the poorest dielectric, the resultant characteristics of the combined insulation of the cable as a whole will be degraded.

The best effect is achieved in the case where the dielectric is removed from the strong greatest-concentration field, and there is an air gap at the center of the cable; this is precisely what happens when a radial air-dielectric combination is used.

With a tangential insulating structure, the dielectric is located along the lines of the electrical field, cutting through the strong concentration. Thus the losses in the dielectric are larger for the tangential combination than for the radial.

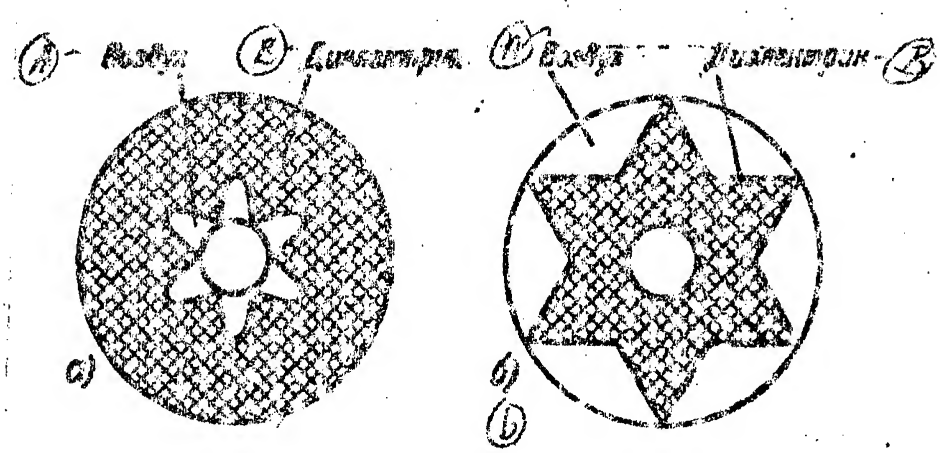


Fig. 5-28. "Star"-Type Insulation. a) Air gap at center; b) air gap at periphery. A) Air; B) dielectric.

When designing coaxial cables with complex multi-layer insulation, it is necessary to combine the dielectric radially, leaving as large an air gap as possible at the center.

From this point of view, comparing the two designs of "star"-type insulation shown in Fig. 5-28, the first is to be preferred, since in this case there is a considerable air gap in the sphere of the maximum effect of the electrical lines of force. On the basis of these considerations, bead-type insulation has rather poor electrical properties in comparison with combined insulation of the continuous type (cord or spiral).

Tables 5-10 and 5-11 give experimentally-obtained

values of ε_e and tan δ_e for various types of cables; it follows from these tables, that from the point of view of the structural form and ratio of volumes, the most common types of dielectric-air insulation may be ranked as follows:

	$\frac{V_d}{V_a}\%$
1) Bead insulation	5
2) Framework of cord and spiral	8-12
3) Caps	13.3

TABLE 5-10

TABLE 5-10. Resultant (equivalent) Values of the Dielectric Constant ε_r for Coaxial Cables of Various Types.

Тип кабеля	Диэлектрик	Тип изоляции	Расстояние между шайбами l, мм	Толщина шайб d, мм	$\varepsilon_э$	Соотношение объемов диэлектрика и воздуха, %
5/18	Керамика	Шайбы	60	3	1,19	5
5/18	Керамика	Шайбы	50	5	1,25	8,3
5/18	Полистирол	Шайбы	60	3	1,03	5
5/18	Стирофлекс	Двухслойная спираль			1,13	12
5/18	Стирофлекс	Опорный каркас			1,13	8
5/18	Полистирол	Колпачки			1,19	13,3
	Бумага	Кордель			1,4—1,6	
	Триацетат	Кордель			1,3—1,5	
	Хлопчатобумажная нить	Спираль			1,12	
1,83/6,7	Полиэтилен	Шайбы	20	1	1,10	5
1,83/6,7	Эбонит	Шайбы	20	1	1,2	5
2,6/9,4	Полиэтилен	Шайбы	25	2,2	1,1	8,8
2,6/9,4	Керамика	Шайбы	25		1,22	
2,6/9,4	Стирофлекс	Опорн. каркас (комбикордель)			1,2	
2,6/9,4	Полистирол	Шайбы	25—30		1,03	
2,6/9,4	Эбонит	Шайбы	20	2,0	1,2	10
2,6/9,4	Полиэтилен	Кордель			1,38	40

A) Type of cable; B) dielectric; C) type of insulation; D) distance between beads b, mm; E) thickness of bead, a, mm; F) $\mathcal{E}_e$; G) ratio of volumes of dielectric and air, %; H) ceramic ; I) polystyrene; J) styroflex; K) paper; L) triacetate; M) cotton paper filament; N) polyethylene; O) ebonite; P) beads; Q) two-layer spiral; R) supporting frame; S) caps; T) cord; U) Spiral; V) supporting frame (combination cord).

As experience in the production and operation of cables has shown, decreasing the relative relationships of the dielectric volumes below the ratios given limits the mechanical stability of the cable insulation.

The various methods of insulation stand in the same order with respect to electrical effectiveness, which for a given dielectric is determined by the ratio of the volume of dielectric V_d, to the volume of air, V_a, in the insulating space of the cable.

The best insulating materials for trunk coaxial cables are polystyrene (styroflex), and polyethylene, which have nearly identical electrical properties in the frequency range of interest to us.

TABLE 5-11

A Тип кабеля	B Диэлектрик	C Тип изоляции
5/18	Керамика F	O Шайбы $a=3$, $б=60$
5/18	Керамика F	Шайбы $a=5$, $б=60$
5/18	Полистирол G	Шайбы $a=3$, $б=60$
5/18	Полистирол G	Колпачки P
5/18	Стирофлекс H	Опорный каркас Q
5/18	Стирофлекс H	Двухслойная спираль R
	Бумага I	Кордель S
	Триацетат J	Кордель
	Хлопчатобумажная нить K	Спираль T
1,83/6,7	Полиэтилен L	O Шайбы $a=1$; $б=20$
1,83/6,7	Эбонит M	Шайбы $a=1$; $б=20$
2,6/9,4	Полиэтилен L	Шайбы $a=2,2$; $б=25$
2,6/9,4	Полистирол G	Шайбы $a=2$; $б=20$
2,6/9,4	Керамика F	Шайбы $a=2$; $б=20$
2,6/9,4	Полиэтилен L	Шайбы $a=2$; $б=20$
2,6/9,4	Полиэтилен L	S Кордель
2,6/9,4	Эбонит M	O Шайбы $a=2$; $б=20$

N П р и м е ч а н и е. Толщина шайбы—a (*мм*). Расстояние между шайбами $б$ (*мм*).

TABLE 5-11
(continued)

D Соотношение $\frac{V_\partial}{V_g}$, %	E $tg\,\delta \cdot 10^{-4}$ при f, гц					
	10^6	$5 \cdot 10^6$	10^7	10^8	$5 \cdot 10^9$	10^{10}
5	1,1	1,0	0,9			
8,3	1,2	1,1	1,1			
5	0,35	0,35	0,3	0,5	0,6	
13,3	0,7	0,7	0,8	1	1,2	
8	0,56	0,6	0,8	0,9	1,0	1,1
12	0,75	0,8	1,0	1,2	1,35	1,5
	130	150				
	40	40				
	7	8				
5	0,37	0,38	0,46	0,7	0,75	
5	14	23				
8,8	0,5	0,5				
10	0,43	0,43	0,43	0,5	0,6	
10	1,76	1,65	1,6			
10	0,6	0,6				
40	1,47	1,47				
10	9,8	20				

TABLE 5-11. Resultant (equivalent) Values of Dielectric Loss Angle Tan δ_e for Coaxial Cables of Various Types. A) Type of cable; B) dielectric; C) type of insulation; D) ratio V_d/V_a, %; E) Tan $\delta \cdot 10^{-4}$ at f, cps; F) ceramic; G) polystyrene; H) styroflex; I) paper; J) triacetate; K) cotton paper filament; L) polyethylene; M) ebonite; N) note. Thickness of bead--a (mm). Distance between beads --b (mm); O) beads; P) caps; Q) supporting frame; R) two-layer spiral; S) cord; T) spiral.

5-9. THE EFFECT OF A DIELECTRIC ON THE CHARACTERISTICS OF A COAXIAL CABLE

a) Ratio of the Metal Attenuation, β_R, to the Dielectric Attenuation, β_G, of a Coaxial Cable

The assumption that $\beta_G = 0$, used in deriving formula (5-31), is correct only in the limiting frequency region. In general, the attenuation of coaxial circuits depends both on the metal loss and the dielectric loss.

In both theory and practice, it has been shown that the properties of the dielectric have a decisive influence upon the parameters of coaxial cables, and consequently, upon the fundamental characteristics of transmission (width of the transmitted frequency band,

quality and span of link).

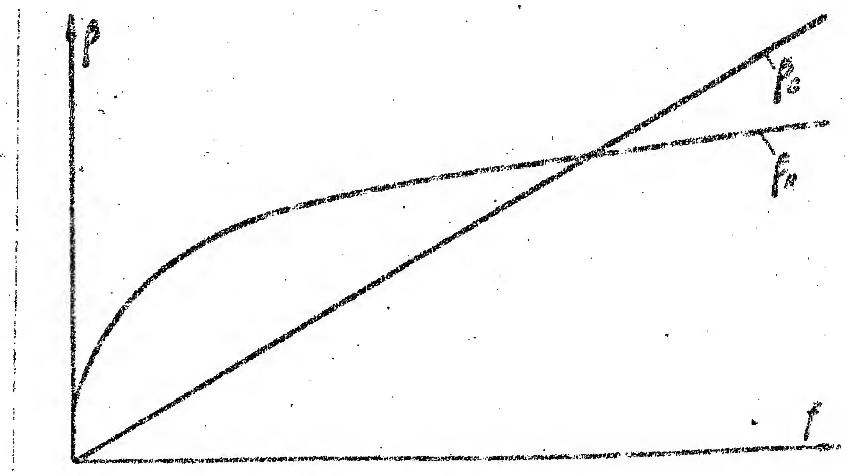


Fig. 5-29. Frequency Dependence of Attenuation Due to Loss in Metal (β_R) and Attenuation Due to Loss in Dielectric (β_G).

This is also supported by the fact that although the theory of coaxial lines was created in 1890, they began to be designed and manufactured only with the appearance of high-frequency dielectrics of "kctop," styroflex, and other types.

Substituting the condition D/d = 3.6 into formula (5-30), the general expression for the attenuation of coaxial circuits may be reduced to the form:

$$\beta = \beta_R + \beta_G = \frac{0{,}25\sqrt{f}\cdot 10^{-3}}{D} + \frac{10}{3}\pi f\sqrt{\varepsilon}\tan\delta\cdot 10^{-6}\ [\text{nepers/km}]$$

(5-52)

In studying the dependence of attenuation upon frequency (5-29), it should be noted that while the first term of equation (5-52) varies in proportion to $\sqrt{f}$, the second term, which is related to the frequency by a linear law, increases considerably faster as f increases.

Thus, if high-quality dielectrics (with low tan δ) are used, it is possible to obtain low dielectric losses in a specific frequency band, so that it may be assumed that $\beta_G \approx 0$; at higher frequencies, however, it will increase so much, that the magnitude of β_G will play a dominant role in the overall attenuation of the cable. Accordingly, let us establish the frequency band for which a coaxial cable is to be used as a function of the type of dielectric employed, introducing the parameter K, which equals the relationship of the attenuation in the dielectric to the attenuation in the metal:

$$K = \frac{\beta_G}{\beta_R} = 0{,}042\sqrt{f}\tan\delta\cdot D.$$

(5-53)

Assuming the overall attenuation of the circuit

to be 1, we obtain an an expression for determining the relative value of β_G and β_R and the total attenuation of the circuit. Since $\beta_R + \beta_G = 1$ and $\frac{\beta_G}{\beta_R} = K$,

$$\beta_G = \frac{K}{K+1} \text{ and } \beta_R = \frac{1}{K+1}.$$

Table 5-12 gives the results of a calculation for the parameter K, and the per cent proportion of attenuation in the dielectric, β_G, and metal, β_R, for a cable D = 9.4 mm, and polyethylene-bead insulation, tan δ = $= 2 \cdot 10^{-4}$. The values of K, β_G, and β_R are also given for the case in which paper-cord insulation is used (tan $\delta \approx 200 \cdot 10^{-4}$).

A) f, гц	B) Диэлектрик	K	β_G, %	β_R, %
10^6	Полиэтилен C)	0,0079	1	99
	Бумага D)	0,788	46	54
10^7	Полиэтилен E)	0,025	2,5	97,5
	Бумага . F)	2,5	71	29
10^8	Полиэтилен G)	0,079	7,3	92,7
	Бумага H)	7,88	89	11
10^{10}	Полиэтилен I)	0,788	44	56
	Бумага . . J)	78,8	98,7	1,3

TABLE 5-12. Frequence Dependence of the Parameters K, β_G, and β_R. A) f, cps; B) dielectric; C) polyethylene; D) paper; E) polyethylene; F) paper; G) polyethylene; H) paper; I) polyethylene; J) paper.

It follows from Table 5-12, that in the frequency region used for modern trunk coaxial cables (up to 10^7 cps), the losses in polyethylene insulation are negligible, and do not exceed 2.5% of the total losses in the cable. For the same frequency range, the losses in paper insulation are 100 times greater than in the polyethylene, and amount to 71% of the total losses. This means that nearly 3/4 of the transmitted electromagnetic energy is lost in the dielectric and, consequently, paper is totally unsuitable for insulating coaxial cables.

The attenuation in the metal depends upon the properties of the materials used, and the geometric dimensions of the cable. The optimum value is obtained by using the ratio $D/d = 3.6$, and by making both conductors of the cable of copper.

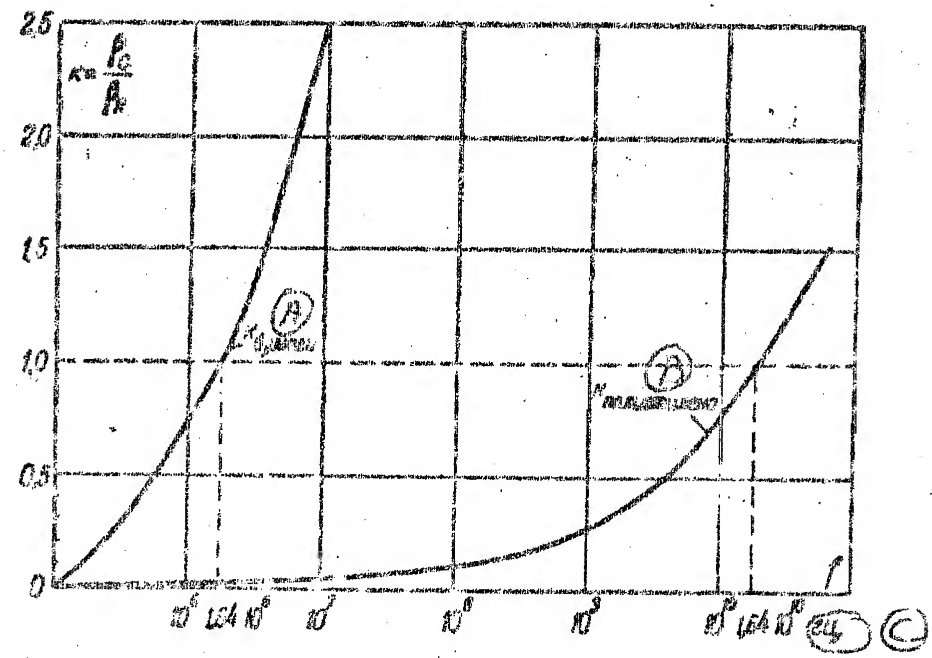


Fig. 5-30. Frequency Dependence of $K = \beta_G/\beta_K$. A) K_{paper}; B) $K_{polyethylene}$; C) cps.

There are no other ways to decrease the attenuation losses in the metal by any substantial amount.

The attenuation owing to the insulation may be reduced to a minimum by improving the dielectric materials used. Thus, while the losses in the metal are unavoidable, the dielectric losses may be attacked successfully.

From the graph of the coefficient K as a function of frequency (Fig. 5-30), it is clear that the attenuations in the dielectric and the metal become equal

$K=\beta_G/\beta_R = 1$) in a cable with air-paper insulation at a frequency of $f = 1.64\cdot10^6$ cps, while in a cable with polyethylene insulation this occurs at $f = 1.64\cdot10^{10}$ cps. Consequently, using polyethylene insulation in cables considerably extends their useful frequency region in comparison with air-paper insulation.

It follows from formula (5-53) that the relative value of attenuation losses in the dielectric increases in proportion to $\tan\delta$ and the diameter D of the cable.

This is explained by the fact that as the diameter of the cable increases, the resistance of the cable drops, and the losses due to attenuation in the metal along with it, and consequently, the relative importance of the attenuation in the dielectric rises.

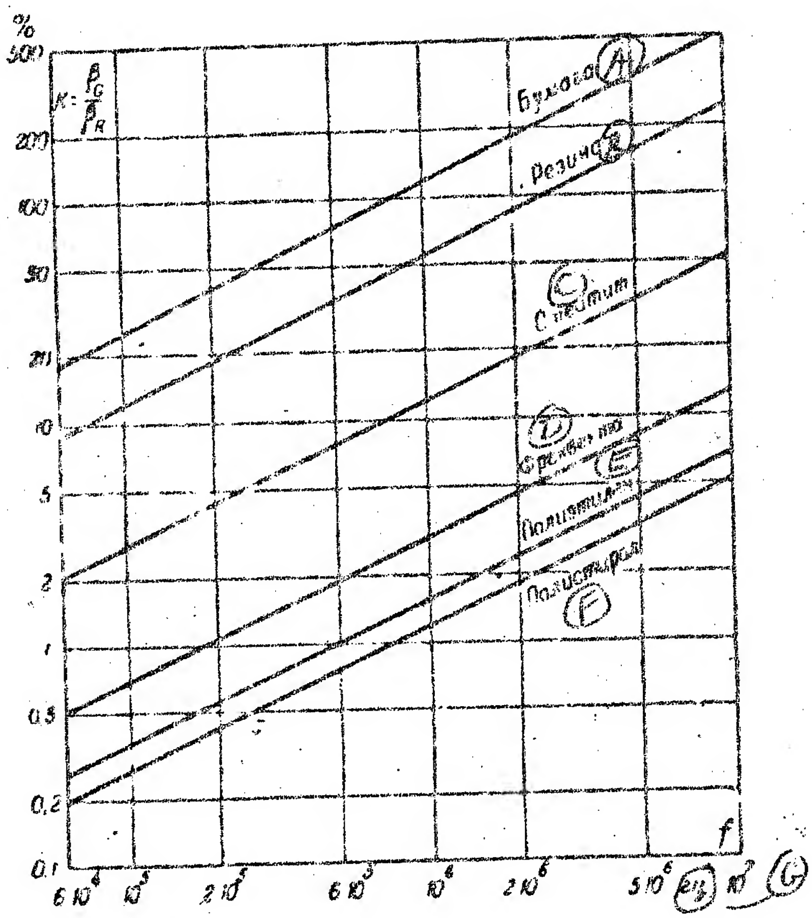


Fig. 5-31. The Ratio of the Losses in the Dielectric to those in the Metal in Coaxial Cables Having Different Dielectrics. A) Paper; B) rubber; C) steatite; D) frekventa; E) polyethylene; F) polystyrene; G) cps.

Fig. 5-31 gives calculated values for the coefficient $K = \beta_G / \beta_R$ (in per cent) for coaxial cables having various dielectrics for the frequency range effectively

transmitted over trunk coaxial cables from $60 \cdot 10^3$ cps to $10 \cdot 10^6$ cps). It follows from Fig. 5-31 that over the entire frequency band up to $10 \cdot 10^6$ cps the losses in such dielectrics as polyethylene and polystyrene do not exceed 7% of the losses in the metal of the cable.

b) Diameter of a Cable With Various Dielectrics

Below we shall consider the effect of the quality of a dielectric upon the dimensions of the cable.

Fig. 5-32 gives the results of a calculation of the diameters of a coaxial cable as a function of the type of dielectrics used, for a frequency of 10^6 cps. The attenuation of all the types of cables used was constant, equalling 0.3 nepers/km.

The outer diameter of a coaxial cable is determined on the basis of formula (5-52), transformed to the following:

$$D = \frac{0{,}25\sqrt{f\varepsilon}\cdot 10^{-8}}{\beta - \frac{10}{3}\pi f\sqrt{\varepsilon}\,\mathrm{tg}\,\delta\cdot 10^{-6}}\ [\textit{мм}]. \tag{5-54}$$

Fig. 5-32 confirms the fact that it is not desi-

rable to use such dielectrics as paper and rubber in coaxial cables. For exactly the same attenuation, the diameter of a polyethylene-insulated cable is 5.3 times less than that of a rubber-insulated cable, and 40 times less than that of a paper-insulated cable. Consequently, it is desirable to use such high-quality dielectrics as polyethylene from the point of view of economizing on the copper and lead used in manufacturing the cable.

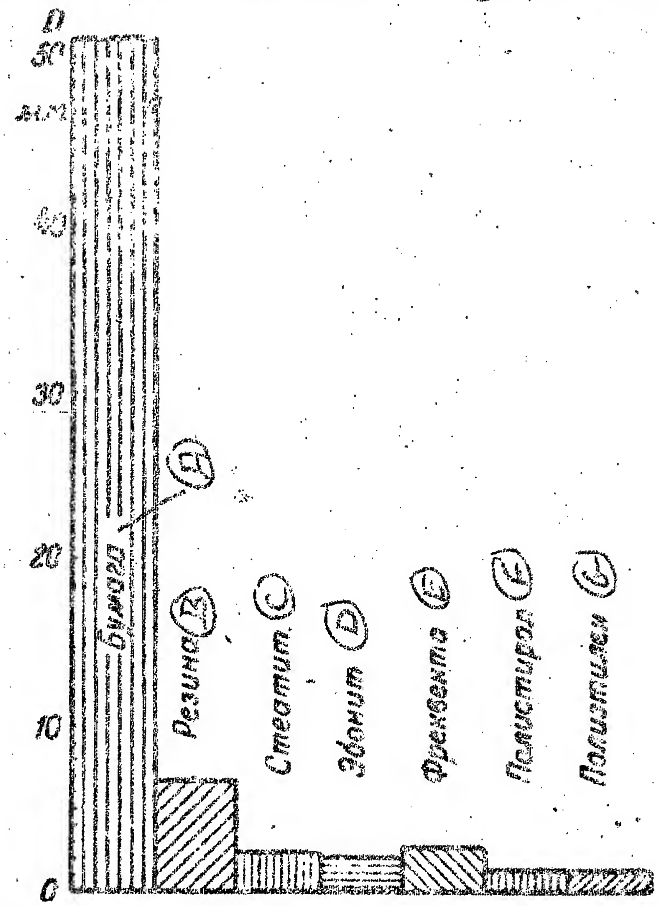


Fig. 5-32. Diameters of Coaxial Cables Using Different Types of Dielectrics ($f = 10^6$ cps). A) Paper; B) rubber; C) steatite; D) ebonite; E) frekventa; F) polystyrene; G) polyethylene.

c) Comparison of the Various Dielectric Designs

In order to illustrate the effect of dielectric structure on the characteristics of a coaxial line, Fig. 5-33 gives the attenuation of a type 2.6/9.4 cable as a function of frequency; this is computed for three types of insulation: 1) Solid polyethylene, 2) polyethylene-bead insulation, 3) air insulation (without a solid dielectric).

It is clear from Fig. 5-33 that even polyethylene introduces additional attenuation into the transmission circuit.

Thus, if the attenuation of an air-insulated cable is assumed to be 100%, bead-type insulation increases the attenuation by 9%, while a solid layer of polyethylene increases it by 54%. This is explained not so much by the losses in the dielectric, β_G, as by the increase in

$\beta_R = (R/2)(\sqrt{C/L})$ owing to the large capacitance of the cable with solid insulation (the capacitance of the cable with solid insulation was C = 99 mμf /km, while for bead-type insulation it was 47.5 mμf/km).

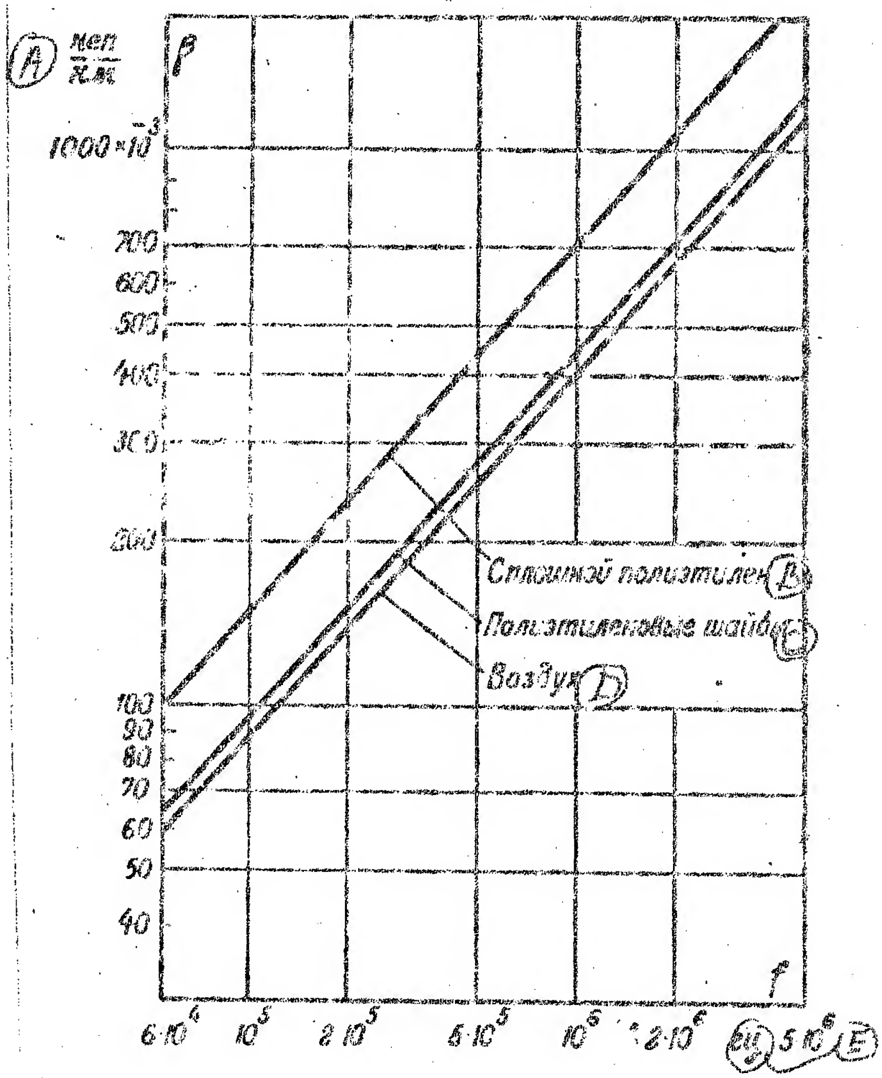


Fig. 5-33. The Attenuation of a type 2.6/9.4 Cable With Polyethylene Insulation As a Function of Frequency. A) Nepers/km; B) solid polyethylene; C) polyethylene beads; D)air; E) cps.

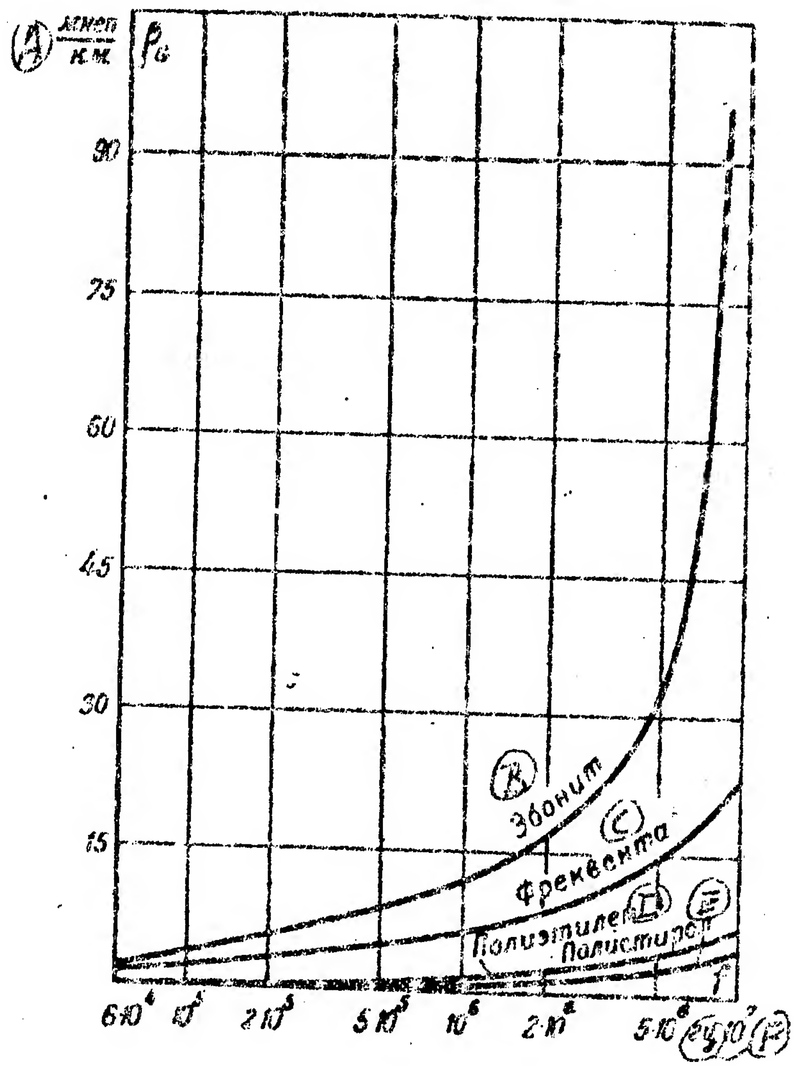


Fig. 5-34. Losses in the Dielectric of Cables Having Various Insulating Beads. A) m nepers/km; B) ebonite; C) frekventa; D) polyethylene; E) polystyrene; F) cps.

For these reasons, as well as to economize upon

materials, in trunk coaxial cables air-combination type insulation is used with the minimum amount of solid dielectric. The amount of the latter used is determined solely by the need for rigid maintenance of the coaxiality of the inner conductor with respect to the outer conductor, and by the requirements for mechanical strength for the cable as a whole.

Fig. 5-34 shows the losses in the dielectric of a 2.6/9.4 type cable having bead-type insulation of ebonite, frekventa, polystyrene, and polyethylene ($a = 2$ mm; $b = 20$ mm); it is clear from the figure that cables with polyethylene and polystyrene insulation have the best electrical properties.

d) Bead-Insulated Cable Design

The bead-type structure is the result of a compromise between two opposing requirements: On the one hand, in order to attain the least attenuation in the cable it is desirable to use the smallest amount of dielectric; on the other hand, the mechanical strength of the cable link and its electrical uniformity require that the beads occur rather frequently in the cable.

Thus b, the distance between beads, and a, the bead thickness, are chosen so as to yield the minimum

amount of dielectric that will simultaneously provide the required mechanical structural strength.

In order to select the bead spacing, it is also necessary to compute the critical wave length. However, trunk cables are utilized for a frequency band lying considerably below the critical wave length, so that in the case considered this factor has practically no importance. For common bead-insulated cables, the critical wave length λ_c is about 10 cm. Consequently, such cables can transmit a frequency band of up to 3,000 Mc.

Fig. 5-35 gives the electrical parameters of type 2.6/9.4 cable with polyethylene bead insulation (bead thickness---2.2 mm) as a function of the variation of the distance between beads from 5 to 50 mm, at a frequency $f = 10^6$ cps.

As Fig. 5-35 shows, increasing the distance between beads decreases the following: The cable capacitance, the shunt conductance, and the attenuation. Increasing b by 10 times (from 5 mm to 50 mm) drops the attenuation by 20%.

Fig. 5-36 shows the dependence of the parameters of the same cable on the variation in the thickness of the polyethylene beads, with a fixed distance between them of b = 25 mm; the figure shows that an increase in a leads

to a decrease in C, G, and β.

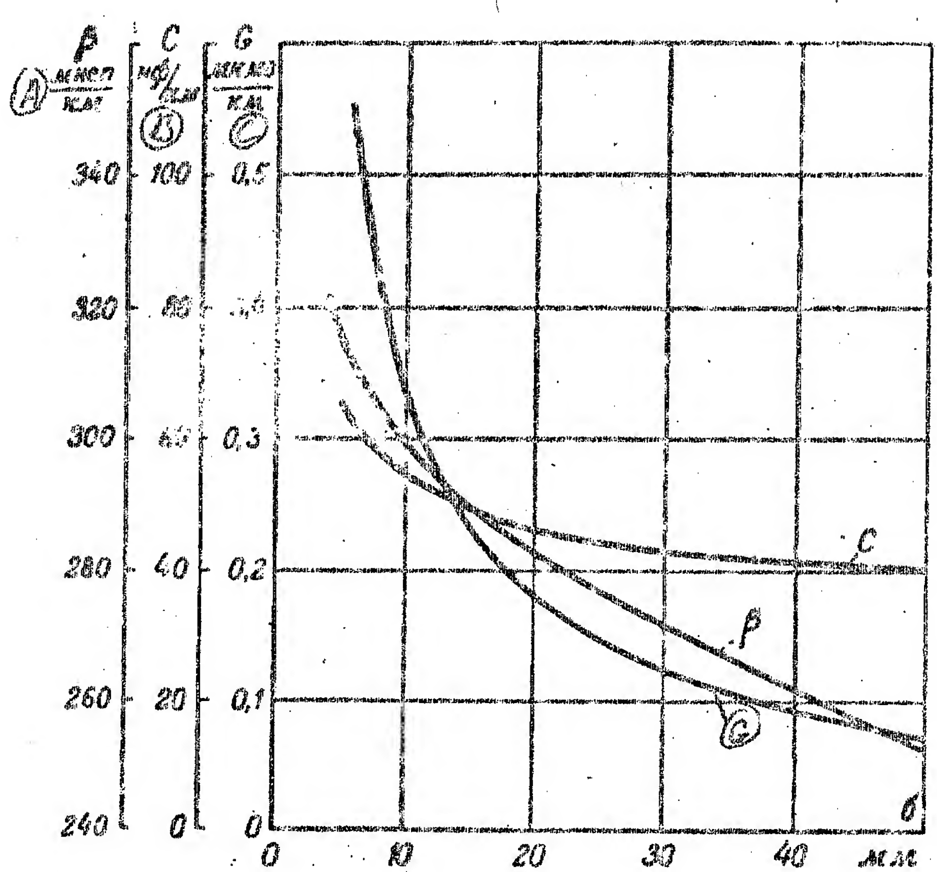


Fig. 5-35. The Dependence of the Parameters of a Cable Upon Variation in the Distance Between Beads ($f = 10^6$ cps). A) m nepers/km; B) $\mu\mu$f/km; C) μmho/km.

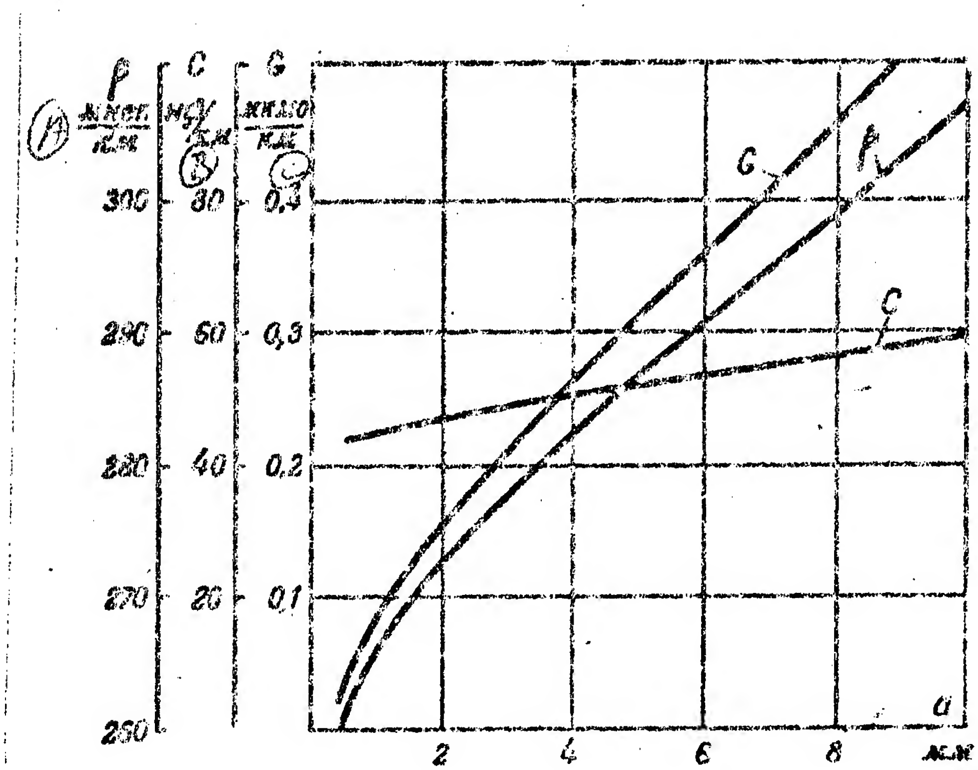


Fig. 5-36. Variations in the Parameters of a Cable With Increasing Thickness of Beads. A) m nepers/km; B) mμf/km; C) μmho/km.

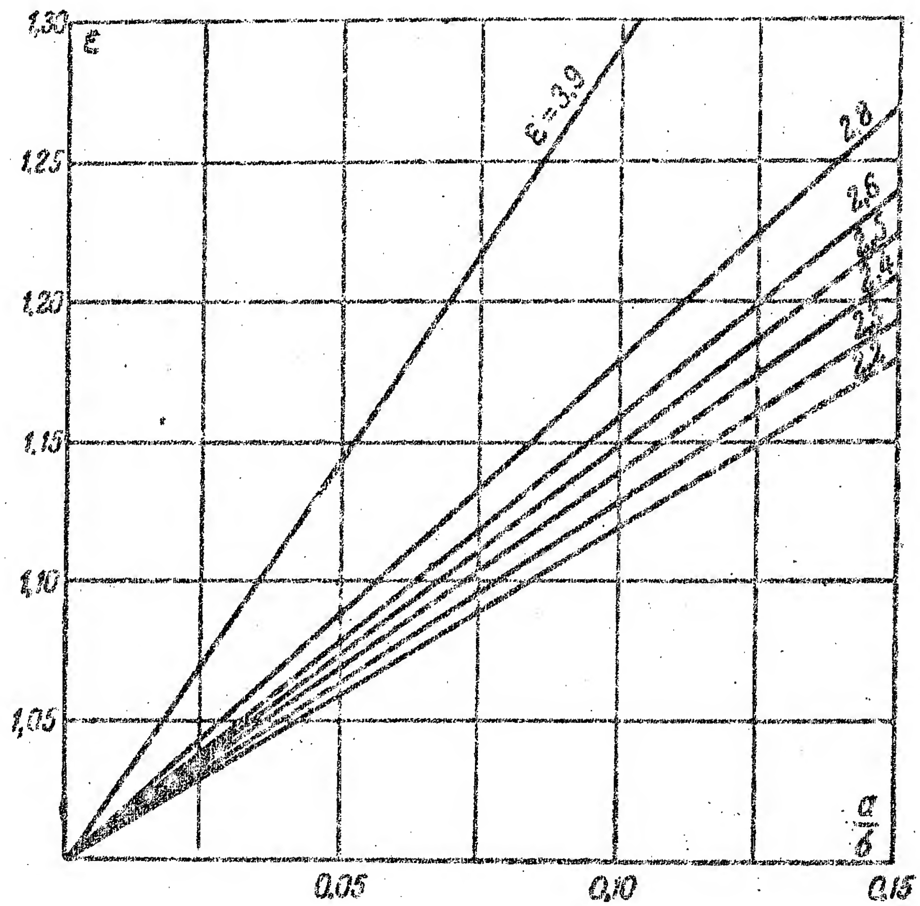


Fig. 5-37. Equivalent Value of Dielectric Constant (ε) As a Function of the Ratio of the Bead Thickness a, and the Bead Separation b for different ε of the dielectric.

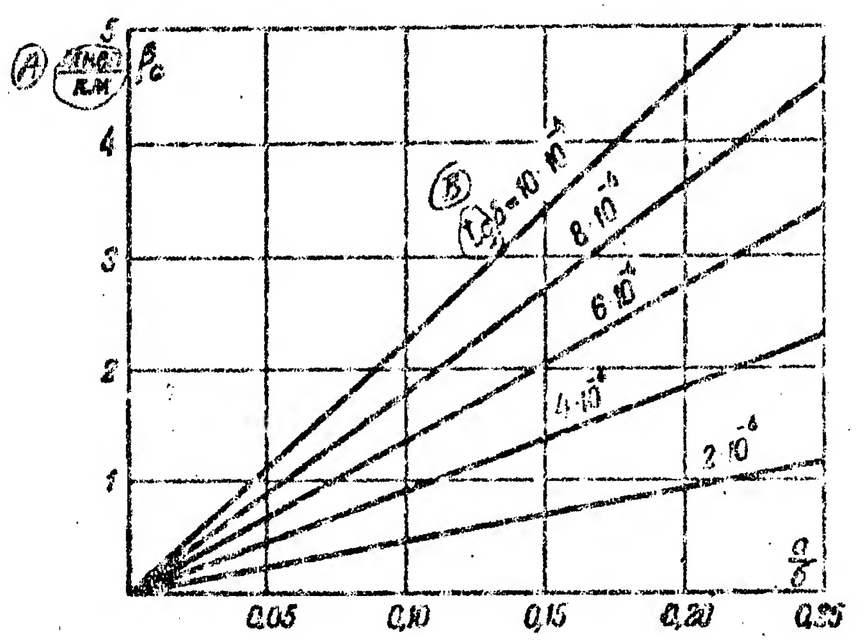


Fig. 5-38. Dependence of the Attenuation Owing to Losses in the Dielectric, β_d, on the Relationship of the Thickness of Beads, a, and the Bead Separation, b, for Various ε of the Dielectric ($f = 10^6$ cps). A) m nepers/km; B) tan.

As the thickness of the beads increases from 0.5 to 10 mm, the cable attenuation rises from 260 to 308 m nepers/km.

On the basis of the requirements for the mechanical strength of a 2.6/9.4 cable, 2.2 mm is chosen as the thickness of the polyethylene beads, and a spacing of 25 mm is used.

Fig. 5-37 gives data which permit the calculation of $\mathcal{E}_e$, for a bead-insulated cable whose dielectric constant, $\mathcal{E}_d$, varies from 2.2 to 3.9. The ratio of the bead's thickness, a, to the bead spacing, b, is plotted along the axis of abscissas, while the value of $\mathcal{E}_e$ is plotted along the axis of ordinates.

Fig. 5-38 gives a graph of the variation of attenuation owing to losses in the bead insulation as a function of the ratio a/b for various dielectrics having a value of tan δ_d from $2 \cdot 10^{-4}$ to $10 \cdot 10^{-4}$, at a frequency $f = 10^6$ cps.

5-10. PRINCIPLES FOR SETTING UP COMMUNICATIONS OVER COAXIAL CABLES

In this chapter, we will consider the following questions that determine the structure of a coaxial cable and the design of a cable trunk as a whole: 1) The principle for employing the coaxial line (2-wire or 4-wire communications systems); 2) link set up (single-cable or double-cable); 3) the principle by which the circuits

are multiplexed (the manner in which telephone and television transmissions are accomplished).

It should be kept in mind that, as a rule, coaxial cables are combination-type. In addition to a fixed number of coaxial circuits, a corresponding number of symmetric pairs and quads are located within a common lead sheath. The symmetric circuits are intended for communications between intermediate sections of the run as well as for signaling and service messages along the trunk.

On the basis for utilizing the long-distance communications circuits, the line is classified as two-wire or four-wire.

In two-wire transmission, a pair of conductors serves for communication in the forward and return direction.

In the four-wire method (Fig. 5-39) four wires are employed for transmission, of which two are used for communications in one direction (from A to B), and the other two--for communications in the return direction (from B to A).

The four-wire communications system is used for trunk long-distance cables using carrier multiplexing. Among the advantages of the system are: Simplicity in

amplifying equipment (differential filters are not required at the repeater points--UP), long transmission span, and reliability of communication.

With the two-wire transmission system, the length of the link, both for symmetric and for coaxial cables, is limited to several hundred kilometers. Accordingly, on all long-distance trunk cables, in order to provide two-way communications, two coaxial circuits are used: One for transmitting all channels in the forward direction, and the other for the reverse direction.

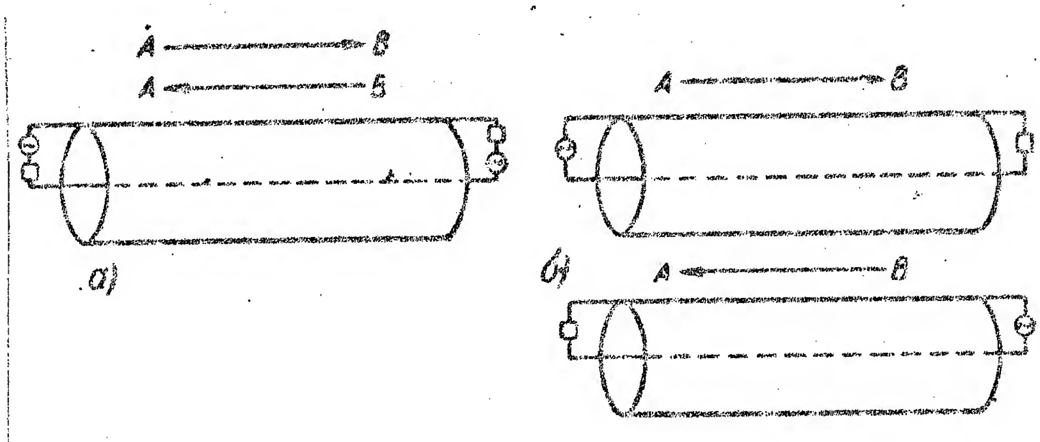


Fig. 5-39. Principles for Using Coaxial Circuits. a) Two-wire principle; b) four-wire principle.

Two systems are used to set up four-wire cable communications: a) The single-cable system, in which all circuits for the forward and return transmission are combined in one cable, and b) the two-cable system, in which the cables of the forward and return links are located in two individual cables.

As a result of many years of experience, the two-cable system has become accepted for symmetric cables.

Two individual cables are laid: in one of them the circuits the A $\longrightarrow$ B direction are grouped, and in the other--the circuits of the B $\longrightarrow$ A direction. The use of the two-cable system has been brought about by the difficulties in protecting symmetrical circuits of direct and return links from mutual interference when they are located in a common cable.

In practical trunk coaxial cable service, both communication systems are used.

In trunk cables using type 5/18 cable, the two-cable communication system is chiefly used. As a rule, the two individual cables are laid in a common trench: One is used for the forward, and the other for the return link.

Medium-size (2.6/9.4) and small (1.83/6.7) cables are used advantageously for single-cable systems. Here,

the presence in these lines of symmetric circuits, whose noise resistance in high-frequency channels (in a single-cable-system) may prove to be inadequate.

Experience in the development and operation of cable trunks has shown that this obstacle may be overcome by the installation of a dividing screen (Fig. 5-41b) or by locating the symmetric circuits in such fashion that the forward links are screened from the return links by the metallic sheathing of the cable (Fig. 5-41a).

The noise resistance of the symmetric circuits may also be increased by limiting the frequency range over which they are used, especially since in a combination cable this is of secondary importance.

All of this points to the greater economy of the single-cable-system of communications, using coaxial trunks.

The methods for multiplexing coaxial trunks have undergone considerable change in recent years.

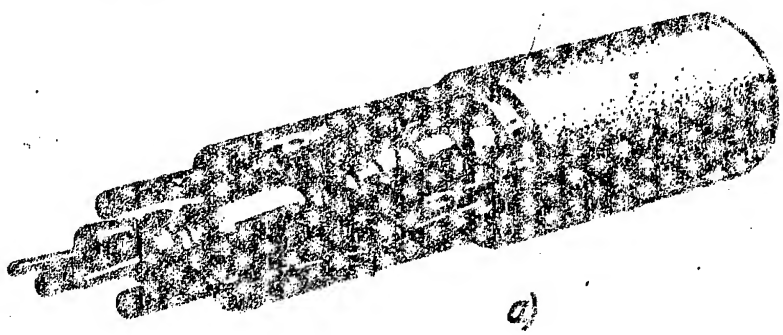
The achievements of contemporary cable technology in the field of shielding coaxial circuits provides complete compatability of the forward and return circuits in one cable.

The only obstruction to the large-scale introduction of single-cable-communications using coaxial lines is

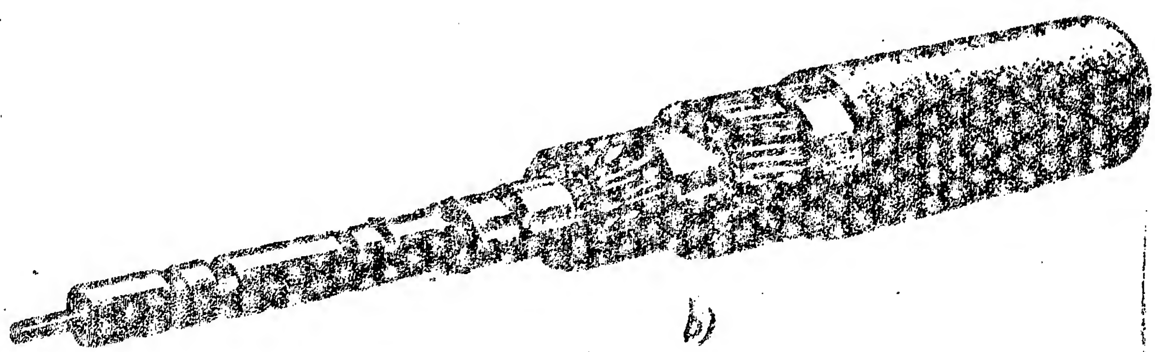
half of the circuits serve transmission in the A $\longrightarrow$ B direction, and the rest for the B $\longrightarrow$ A direction.

Fig. 5-40 shows combination coaxial cables used in single-cable and double-cable systems.

In comparing these communications systems, it should be noted that the use of two separate cables for communication in the forward and return direction is uneconomical, and is the first stage in the introduction of coaxial lines.



a)



b)

Fig. 5-40. a) Cable for single-cable-system communication; b) cable for double-cable-system communication.

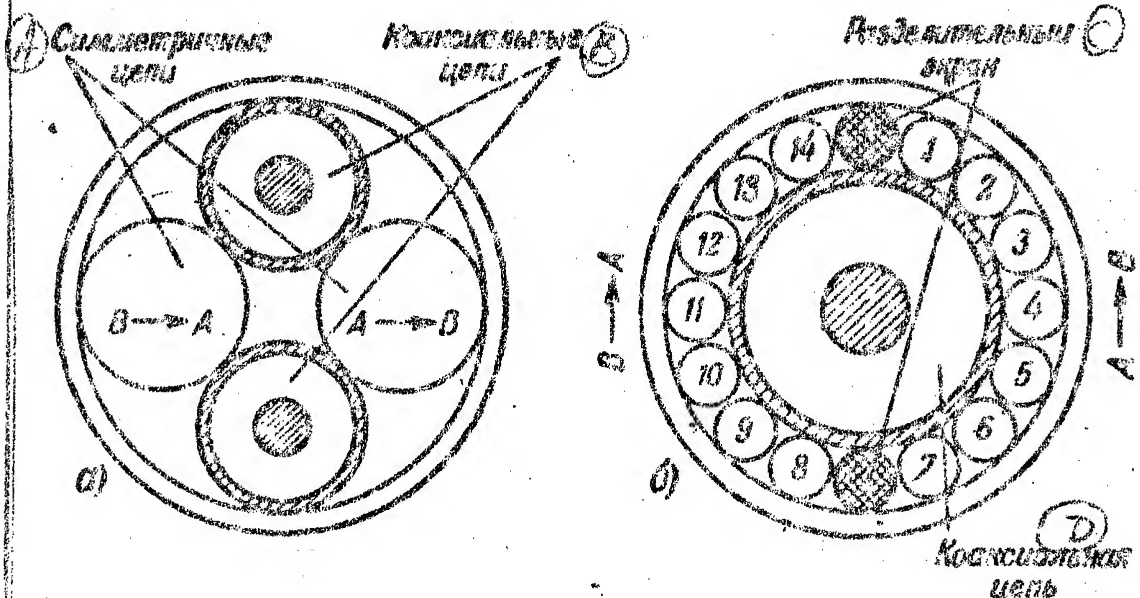


Fig. 5-41. Shielding of Symmetric Circuits in a Combination Cable. 1-7--A ——→ B symmetric circuits; 8-14--B ——→ A symmetric circuits. A) Symmetric circuits B) coaxial circuits; C) dividing shield; D) coaxial circuit.

At first, when the number of telephone links existing in a coaxial cable did not exceed 100-200 (up to 1-2 Mc) and 2-3 Mc were reserved for television, the multiplexing spectrum did not exceed 4-5 Mc, and the entire transmission utilized a single cable. This, in particular,

was the multiplexing system for type 5/18 and 1.83/6.7 cables.

At present there is a tendency to multiplex coaxial circuits by using separate coaxial pairs for the transmission of telephone and telegraph channels. For example, in type 2.6/9.4 coaxial cables, two coaxial circuits are reserved for television (forward and return transmission) and two circuits for two-way telephone communication.

The separation of coaxial circuits into television and telephone circuits was carried out on the basis of the following considerations: 1) Modern systems of high-quality television require a band width of up to 6-10 Mc.

In modern trunk links the number of telephone channels reaches 660, using a frequency band of 3 Mc.

Thus, in order to combine telephone links and television in one circuit it would be necessary to use the coaxial cable over a 9-13 Mc frequency spectrum, which would lead to a corresponding increase in the communications span (distance between repeater points), 2) the equipment required to separate the individual telephone and television circuits is simplified (dividing equipment is not required at the UP). The distortion compensation system is also simplified, since the different channels require different equalizing circuits (in tele-

vision it is primarily important to remove phase distortion, while for telephone circuits amplitude distortion must be eliminated).

What has been said above makes it possible to formulate the basic principles for constructing coaxial cables and designing modern cable trunks.

1. Coaxial trunks should be used in a single-cable system, which is more economical and technically progressive.

2. All links should be set up in a four-wire system, with separate coaxial circuits for television and telephony.

3. The most advantageous construction of a combined coaxial cable is the location under a common lead sheath of four coaxial pairs and a corresponding number of symmetric circuits (used in multiplexing up to 60 kc, and in low-frequency communications service).

5-11. THE STRUCTURE OF COMBINATION COAXIAL CABLES

There are several types of combination coaxial cables. The most typical of them are the following.

TYPE 2.6/9.4 COAXIAL CABLE (Fig. 5-42)

The cable consists of four coaxial pairs, located at the center, and one layer of symmetric circuits, containing two shielded pairs, and 10-14 spiral quads for carrier communication. The shielded pairs are located diametrically opposite each other. In the empty space between the coaxial pairs, there are located four quads intended for service communication (low-frequency communication).

Each coaxial pair consists of an inner copper conductor, d =2.6 mm, and an outer conductor which takes the form of a copper tube with a diameter of D =9.4 mm; the tube is the single-seam "zipper " type.

The coaxial pairs are insulated with polyethylene beads, 2.2 mm thick, spaced 25 mm apart. The shield around the outer conductor consists of two mild-steel tapes, 0.15-0.2 mm thick.

The four coaxial pairs and the four symmetric service quads are combined into a strand, covered by a winding of 2-3 layers of paper tape. The diameter of of the symmetric circuits of the spiral quad is 1.2 mm, the diameter of the shielded pair is 1.4 mm. Paper-cord insulation is used. The shield is made of metalized paper.

One of the shielded pairs is for monitoring, and its conductors have enamel insulation.

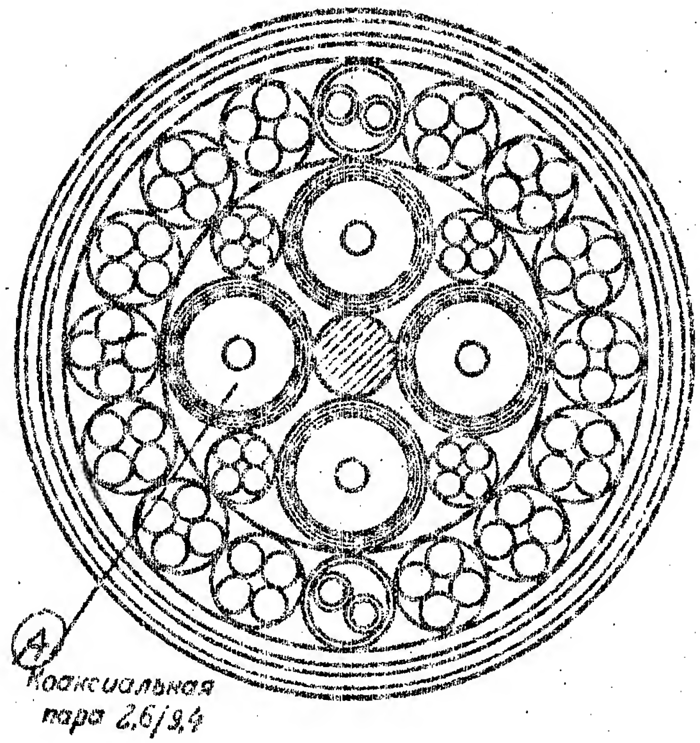


Fig. 5-42. Type 2.6/9.4 combination cable.
A) 2.6/9.4 coaxial pair.

Each symmetric group is covered with paper tape. The twisted cable is covered with 2-3 layers of paper and placed in a lead sheath. The surface of the zinc sheath is covered with 2-3 layers of paper and a layer of hemp impregnated with an asphalt compound.

The cable is armored with two steel tapes, 0.5 mm

thick each. A layer of hemp, impregnated with a bituminous compound and bearing a chalk solution, covered the surface of the armoring. The factory shipping length of the cable is 425 m.

Table 5-5 gives the electrical parameters of the 2.6/9.4 coaxial pairs. An over-all view of the cable is given in Fig. 5-43.

Two diametrically opposite coaxial pairs serve for setting up 660 communications channels with a frequency ranging from 60 cps to 3000 cps. The cable is used in the four-wire communications system. The other two coaxial pairs are intended for television transmission and have a pass-band of from 60 cps to $8 \cdot 10^6$ cps. The repeater stations are 9-12 mm apart.*

* As in original — Translator's note.

The high frequency balanced quads are multiplexed, using frequencies ranging up to 60,000 cps; they serve to set up a twelve channel link between intermediate points along the trunk.

Type 5/18 Combination Coaxial Cable
(Figs. 5-40 and 5-44)

The cable consists of a single coaxial pair, located at the center, and one layer of 26 symmetric circuits

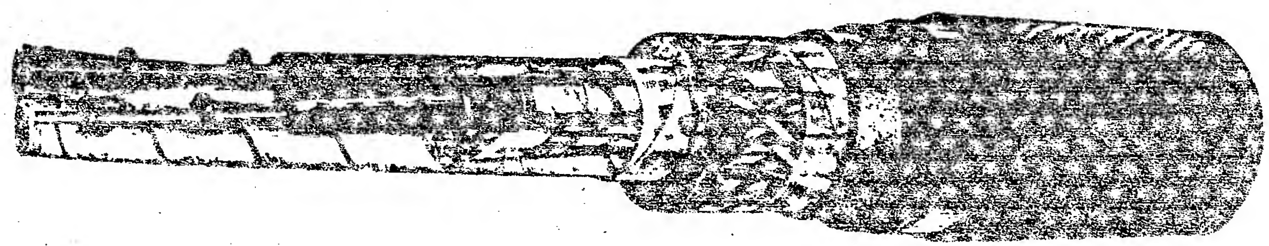


Fig. 5-43. Over-all view of type 2.6/9.4 combination coaxial cable.

Fig. 5-44 shows the structure of a 27-d cable (the number 27 shows the total number of pairs in the cable).

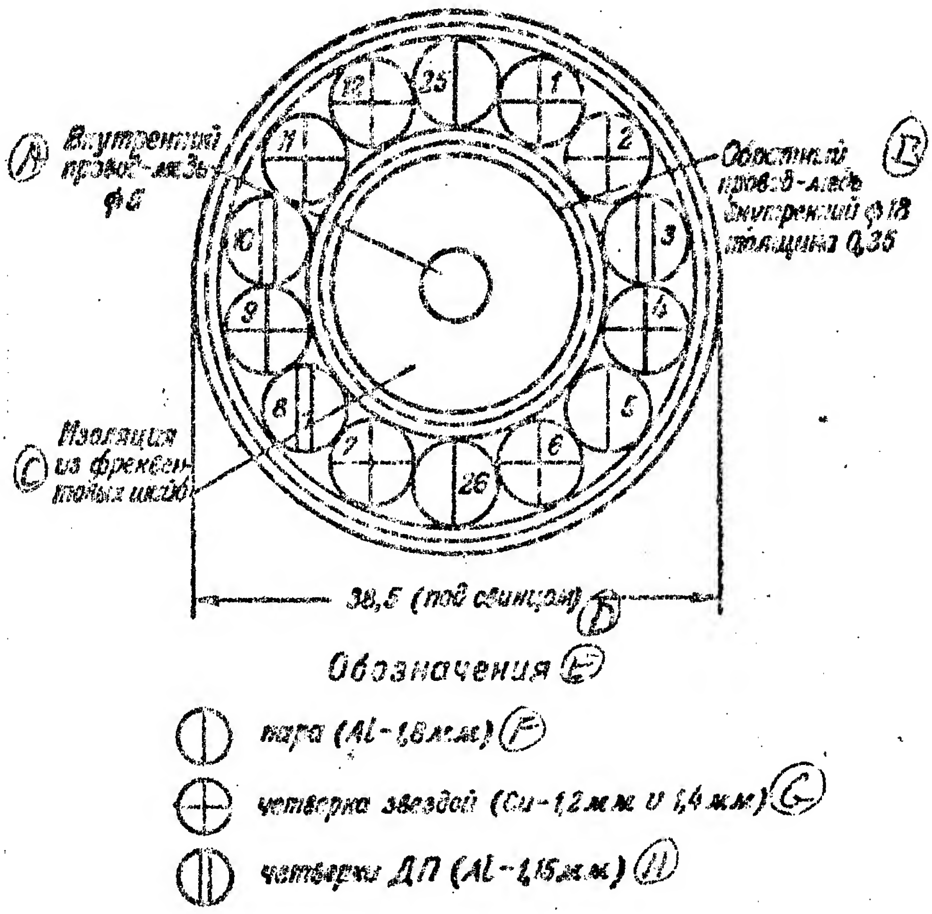


Fig. 5-44. The 27-d type (5/18) combination coaxial cable. A) Inner conductor — copper diameter 5; B) return conductor — copper inside diameter 18, thickness 0.35; C) "frekventa-bead insulation; D) 38.5 (under the lead); E) symbols; F) pair (Al — 1.8 mm); G) spiral quad (Cu — 1.2 mm and 1.4 mm); H) double pair quad (Al — 1.15 mm).

The dimensional ratio of the coaxial pair is d/D =

= 5/18. The conductors are copper. The insulation is made up of frekventa beads, 3 mm thick, spaced 60 mm apart. The outer conductor is of the tubular type. Two pairs and twelve quads are located in the layer.

For radiobroadcasting, pairs with aluminum conductors, 1.8 mm in diameter, are used. Six spiral quads with copper conductors, 1.2 mm in diameter (Nos. 1, 2, 6, 7, 11, and 12) are used for twelve channel carrier multiplexing with a frequency spectrum of up to 60,000 cps. Two similar quads with copper conductors 1.4 mm in diameter (Nos. 4 and 9), in the coil-loaded form, are used in a three-channel multiplexed system. Four two-pair quads have aluminum conductors, 1.15 mm in diameter (Nos. 3, 5, 8, and 10), are lightly coil-loaded, and carry a single carrier channel.

The shielding and armoring of the cable is standard. The outside diameter is 52 mm. The cable weighs 6,600 kg per km. The factory shipping length is 425 m.

The combining of a large number of groups differing in structure into a single cable makes it possible to use various types of multiplexing equipment; this increases the flexibility of the cable in service.

The transmission span for the various systems of communications used in this combination cable differ (17.5 km, 35 km, 70 km, and 140 km).

Table 5-13

A Тип кабеля	B Внутренний провод	C Изоляция	D Внешний провод — E Конструкция
27а	5 *мм* — медь K	Шайбы из фреквенты M	Медная труба O $\Delta = 0{,}35$
27б	5 *мм* — медь K	Каркас из стирофлексных спиралей и лент N	Медные полутрубки $\Delta = 0{,}35$ P
27д	5 *мм* — медь K	Шайбы из фреквенты M	Медная труба O $\Delta = 0{,}35$
27е	5 *мм* — медь K	Каркас из стирофлексных спиралей и лент N	Медные полутрубки $\Delta = 0{,}35$ P
27д	4,6 — алюминий L +0,2 — медь K	Шайбы из фреквенты M	Алюминиевая труба Q $\Delta = 0{,}6$
27ж	4,5 — алюминий L +0,25 — медь K	Каркас из стирофлексных спиралей и лент N	Алюминиевые полутрубки $\Delta = 0{,}5$ R

Table 5-13
Continued

Внутренний диаметр, мм	Толщина свинцовой оболочки, мм	Диаметр под свинцом, мм	Броневой покров	Наружный диаметр кабеля, мм
18	2,3	38,5	Плоская проволока и джут	52
18	2,3	38,5	То же	55
18	2,3	38,5	»	52
18	2,3	38,5	»	55
19,5	1,5—2,3	40,0	»	55
19,5	1,5—2,3	40,0	»	54

Table 5-14

(A) Тип кабеля	(B) Цепи радиовещания	(C) Непупинизированные высокочастотные цепи			
		(D) Конструкция четверки	(E) Тип скрутки	(F) Число связей	(G) Дальность связи, км
27-а	2×2—1,8 мм—А	6×4—1,2 мм—М	3	12	35
27-б	2×2—1,8 мм—А	6×4—1,2 мм—М	3	12	35
27-д	2×2—1,8—А	6×4—1,2—М	3	12	35
27-е	2×2—1,8—А	6×4—1,2—М	3	12	35
27-ж	2×2—1,8—А	6×4—1,2—М	3	12	35
27-з	2×2—1,8—А	6×4—1,2—М	3	12	35

(Q) Примечание. А—алюминий, М—медь.

Table 5-14
Continued

Конструкция четверки I	Тип скрутки J	Система пупинизации K	Число связей L	Дальность связи, км M
H Пупинизированные цепи				
6×4—1,15 мм—А	ДП	Средняя N	1	140
6×4—1,15 мм—А	ДП	Средняя N	1	140
4×4—1,15—А	ДП	Средняя N	1	140
2×4—1,4—М	З	Очень легкая O	3	70
4×4—1,15—А	ДП	Средняя N	1	140
2×4—1,4—М	З	Очень легкая O	3	70
4×4—1,15—А	ДП	Легкая P	2	70
2×4—1,4—М	З	Очень легкая O	3	70
4×4—1,15—А	ДП	Легкая P	2	70
2×4—1,4—М	З	Очень легкая O	3	70

Table 5-13. Structural Data for Large Combination Cable, Type 5/18. A) Type of cable; B) inner conductor; C) insulation; D) outer conductor; E) structure; F) inside diameter, mm; G) thickness of lead sheath, mm; H) diameter under lead, mm; I) armoring; J) outside diameter of cable, mm; K) copper; L) aluminum; M) frekventa beads; N) frame of styroflex spiral and tape; O) copper tube; P) copper half tube; Q) aluminum tube; R) aluminum half tube; S) flat wire and jute; T) the same.

Table 5-14. Basic Characteristics of the Symmetric Circuit for Large Type (5/18) Combination Cables. A) Type of cable; B) radiobroadcasting circuit; C) non-coil-loaded high frequency circuits; D) quad structure; E) type of lay; F) number of links; G) communications span, km; H) coil-loaded circuits; I) quad structure; J) type of lay-up; K) coil-loading system; L) number of links; M) communications span, km; N) medium; O) very light; P) light; Q) Note. A — aluminum, M — copper.

In addition to the 27-d type cable, cables designated 27-a, 27-b, 27-e, 27-f, and 27-zh are in use. They represent structural modifications and are used in the same fashion as the 27-d cable. Cables 5-13 and 5-14 give the structural data for coaxial and symmetric circuits of cables of these types.

In multiplexing the 5/18 coaxial pair, a 90 to 690 kc bandwidth is used to obtain 200 telephone channels and a $1 \cdot 10^6$ to $4 \cdot 10^6$ cps bandwidth for television transmission. The distance between repeater points for the telephone link is 35 km, and for the television link, 17.5 km. Two separate combination cables are laid for the forward and return directions of transmission; they follow the same route.

There is a type of 5/18 coaxial combination cable which combines two coaxial pairs and a corresponding number of symmetric circuits under a common lead sheath. With such a cable, it is possible to set up a single-cable system of communications (Fig. 5-40a).

Type 1.83/6.7 Combination Coaxial Cable

Type 1.83/6.7 combination cables exist having two, four, six, and even eight coaxial pairs within a single lead sheath.

The structure of a four-coaxial combination cable

with four coaxial pairs is as follows: the dimensional ratio of the coaxial pair is 1.83/6.7.

The insulation consists of ebonite or polyethylene beads, 1.6 and 1.78 mm thick, respectively. The beads are about 20 mm apart.

The outer conductor has a single "zipper"-type seam. The coaxial pair is shielded with two steel tapes.

In the center of the cable there is a symmetric quad for signalling, having copper conductors 0.64 mm in diameter (for signalling). In the empty space between the coaxial pairs there are four combination quads (for service communications along the cable run), in which two of the diagonally located pairs have conductors 0.91 mm in diameter, and the other two have 0.64 mm diameter conductors. In the outer layer there are 18 symmetric quads with 0.91 mm diameter conductors. These circuits carry a 12-channel system of carrier telephony. The symmetric circuits are air-paper insulated. Two diametrically opposite coaxial pairs serve to carry 480 telephone conversations with a frequency range of from 64 to 2064 cps. The other two pairs are used for television transmission.

The distance between repeaters is 8.5 km. Attended repeater points are set up every 80-100 km.

Fig. 5-45 shows a combination coaxial cable con-

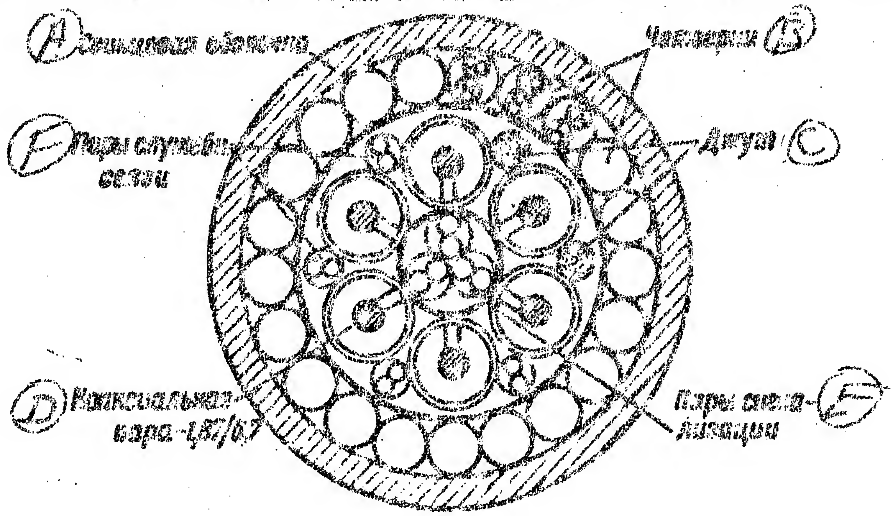


Fig. 5-45. Type 1.83/6.7 small coaxial combination cable. A) Lead sheath; B) quads; C) jute; D) coaxial pairs – 1.87/6.7; E) signalling pair; F) service-communications pair.

sisting of six coaxial pairs and a corresponding number of symmetric quads. The structure and technical data of basic types of combination cables are given in Tables 5-15 and 5-16.

5-12. NONUNIFORMITIES IN COAXIAL CABLES

Owing to structural and production variations, the dimensions of the conductors and dielectric of the cable are not entirely uniform along the length of the cable. These internal nonuniformities have an effect upon the parameters of the cable, since the coaxial circuit ceases to be uniform along its entire length. The effectly chiefly shows up in wave impedance of the cable, the magnitude of which differs from the nominal value in sections containing nonuniformities.

TABLE 5-15.

Structural data for trunk coaxial cables. A) type of cables; B) inner conductor, mm; C) copper; D) outer conductor; E) construction; F) inside diameter, mm; G) "Molniya" ("zipper") copper; H) overlapping copper half tubes; I) copper, 12 Z-shaped strips; J) insulation; K) polyethylene beads (A = 2.2 mm); L) shell formed of styroflex spirals and cords; M) beads of ebonite and polyethylene (1.6--1.78 mm); N) kotopa cord; O) super-kotopa cord and ebonite beads; P) number of coaxial pairs; Q) number of symmetric groups; R) 26 symmetric pairs.

TABLE 5-15

A Тип кабеля	B Внутренний провод, *мм*	D Наружный провод: E Конструкция	D Наружный провод: F Внутренний диаметр, *мм*	J Изоляция	P Число коаксиальных пар
2,6/9,4	C Медь 2,6 *мм*	Медь „Молния" G	9,4	Полиэтиленовые шайбы (a=2,2 *мм*) K	4
5/18	C Медь—5	Медные полутрубки с перехватами H	18	Каркас из стирофлексных спиралей и корделей L	1
1,83/6,7	C Медь—1,83	Медь „Молния" G	6,7	Шайбы из эбонита и полиэтилена (1,6—1,78 *мм*) M	4—6
3,17/11,7	C Медь—3,17	Медь, 12 Z-образных лент I	11,7	Кордель из котопы N	4
2,64/9,5	C Медь—2,64	То же I	9,5	Кордель из супер-котопы и шайбы из эбонита O	2

Table 5-15

Количество симметричных групп
14×4—1,2 мм+2×2—1,4 мм+4×4—1,2 мм
25 симметричных пар (d=0,9; 1,2 и 1,4) and
13×4—0,91 мм+4×4—0,64 мм и 0,91 мм
6×4+4×2 (d=0,91 и and 1,3 мм)
20×4 (d=0,91 и 1,3 мм) and

TABLE 5 16.

Basic technical data and multiplexing systems for trunk-line coaxial cables. A) type of cable; B) telephony; C) number of channels; D) frequency range, Mc; E) up to; F) television range, Mc; G) up to 6.0 (black and white); up to 8.0 (color); H) distance between repeater points, km; I) 35 telephone; 17.5 television; J) communications system; K) single cable; L) two-cable; M) wave impedance, ohms; N) capacitance, millimicro farads/km; O) attenuation at 1 Mc, nepers/km; P) television circuits; Q) separate; R) combined with telephone circuits.

(A) Тип кабеля	(B) Телефонирование: (C) Количество каналов	(D) Диапазон частот, Мгц
2,6/9,4	660	(E) До 3,0
5/18	200	До 0,69
1,83/6,7	240	До 1,0
	480	До 2,06
3,17/11,7	160	До 1,0
2,64/9,5	630	До 2,79
2,5/9,0	180	До 0,788

TABLE 5-16.

F Диапазон телевидения, *мггц*	H Расстояние между усилительными пунктами, *км*	J Система связи	M Волновое сопротивление, *ом*	N Емкость, *нф/км*	O Затухание при 1 *мггц*, *неп/км*	P Цепи для телевидения
G До 6,0 (черное — —белое) До 8,0 (цветное)	9—10	K Однокабельная	75	48	0,26	Отдельные Q
1—4	I 35 телефон. 17,5 телевиден.	Двухкабельн. L	70	50	0,16	Совмещены с телефонными R
1—5	17	K Однакабельная	75	47	0,40	Совмещены с телефонными R
E До 5—8	8,5	K Однокабельная	75	47	0,40	Отдельные Q
E До 5	17	K Однокабельная	70	50	—	Отдельные Q
—	9,6	K Однокабельная	70—75	90	0,30	—
E До 3,0		K Однокабельная				—

As a result, as an electromagnetic wave propagates along the cable, it encounters a nonuniformity in its path, is partially reflected by it, and a portion of it is returned to the beginning of the line. When there are several nonuniform sections, the wave undergoes a series of partial reflections, and, circulating along the line causes additional attenuation and distortion of the circuit characteristics.

The nonuniformities in the cable lead to the appearance of two further energy flows: 1) a flow of reflected energy, consisting of the sum of the elementary reflections at the sites of nonuniformities, moving toward the beginning of the line; 2) a forward energy flow, arising on account of double reflections and moving toward the end of the line together with the basic energy transmitted along the cable. The forward current arises as the original reflected waves, moving toward the beginning of the line, encounter points of nonuniformity and are partially reflected toward the end of the line.

The flow of reflected energy causes variations in the input impedance of the cable, z_{in}. The magnitude of the input impedance varies and ripple occurs. This makes it difficult to match the cable to the equipment at the ends of the line and leads to distortion in

the transmission circuit.

The forward current, propagated together with the basic current, arrives at the receiver, distorts the shape of the transmitted signal, and also causes noise in transmission. It has an especially unfavorable effect on the quality of a television transmission, where the phase relationship of the transmitted and received signals is a critical factor.

It has been established experimentally that in order to carry out normal transmission of television signals, the value of the forward current must not amount to more than 1% of the basic current.

A very important requirement for television is the absence of amplitude distortion in the transmission circuit, and it is of primary importance to try to keep z_{in} constant.

Figure 5-46 gives the typical frequency dependence of the input impedance z_{in} of a type 5/18 coaxial cable. The value of z_{in} fluctuates with respect to the value of the wave impedance of the cable, z.

The wave impedance of a coaxial cable is determined in accordance with formula (5-15), and consequently depends upon the three factors b, D, and $\mathcal{E}$.

Keeping in mind that the nonuniformity of these

values, Δd, ΔD, and $\Delta \varepsilon$, is comparatively slight, the deviation of the wave impedance from the average value (ripple) may be expressed by the formula

$$\Delta z = \frac{60}{\sqrt{\varepsilon}}\left(\frac{\Delta D}{D} - \frac{\Delta d}{d} - \frac{\Delta \varepsilon}{2\varepsilon}\ln\frac{D}{d}\right) \text{(ohm)}, \quad 5\text{-}55$$

where

$\frac{60}{\sqrt{\varepsilon}}\frac{\Delta D}{D} = \Delta z_D$ is the deviation of z due to nonuniformity of the outer conductor;

$-\frac{60}{\sqrt{\varepsilon}}\frac{\Delta d}{d} = \Delta z_d$ - is the deviation of z owing to nonuniformity of the inner conductor;

$-\frac{60}{\sqrt{\varepsilon}}\frac{\Delta \varepsilon}{2\varepsilon}\ln\frac{D}{d} = \Delta z_\varepsilon$ - is the deviation of z owing to nonuniformity in dielectric constant.

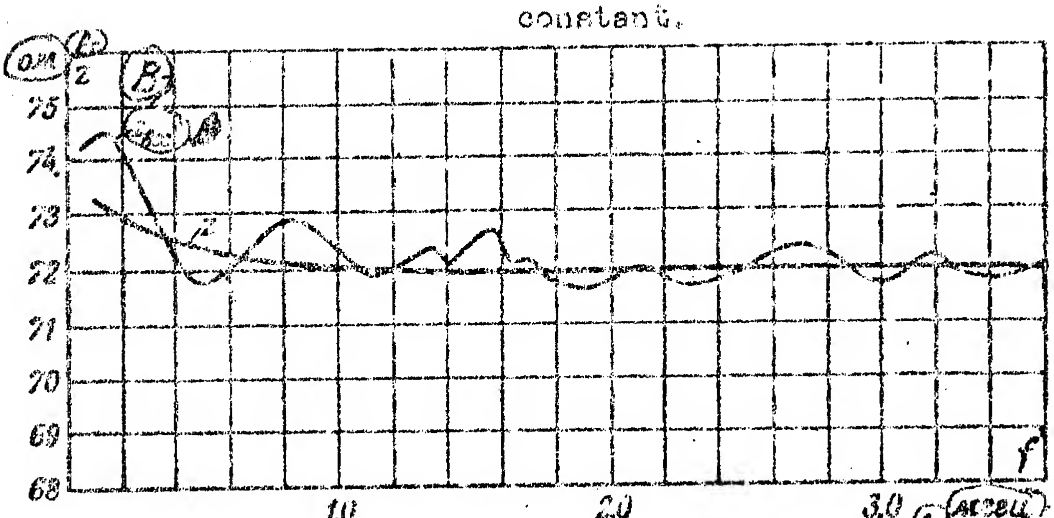


Fig. 5-46. Input impedance of a

coaxial cable as a function of frequency. A) ohms; B) z_{in}; C) Mc.

In practice, the ripple Δz is assumed to be determined not in absolute unity, but relative to the value of the wave impedance of the cable, , i.e.,

$$\frac{\Delta z_D}{z} = \frac{\Delta D}{D} \cdot \frac{1}{\ln \frac{D}{d}}; \quad \frac{\Delta z_d}{z} = -\frac{\Delta d}{d} \cdot \frac{1}{\ln \frac{D}{d}}; \quad \frac{\Delta z_\varepsilon}{z} = -\frac{\Delta \varepsilon}{2\varepsilon}.$$

Investigations have shown that the greatest problems are connected with variation in the diameter and thickness of the outer conductor. A considerable part is also played by nonuniformity along the cable owing to all the possible indentations, overlaps, and other departures from the shape of an ideal hollow cylinder caused by manufacturing conditions and the requirements for cable flexibility.

The inner conductor of the cable is manufactured fairly accurately, and contributes little to the ripple.

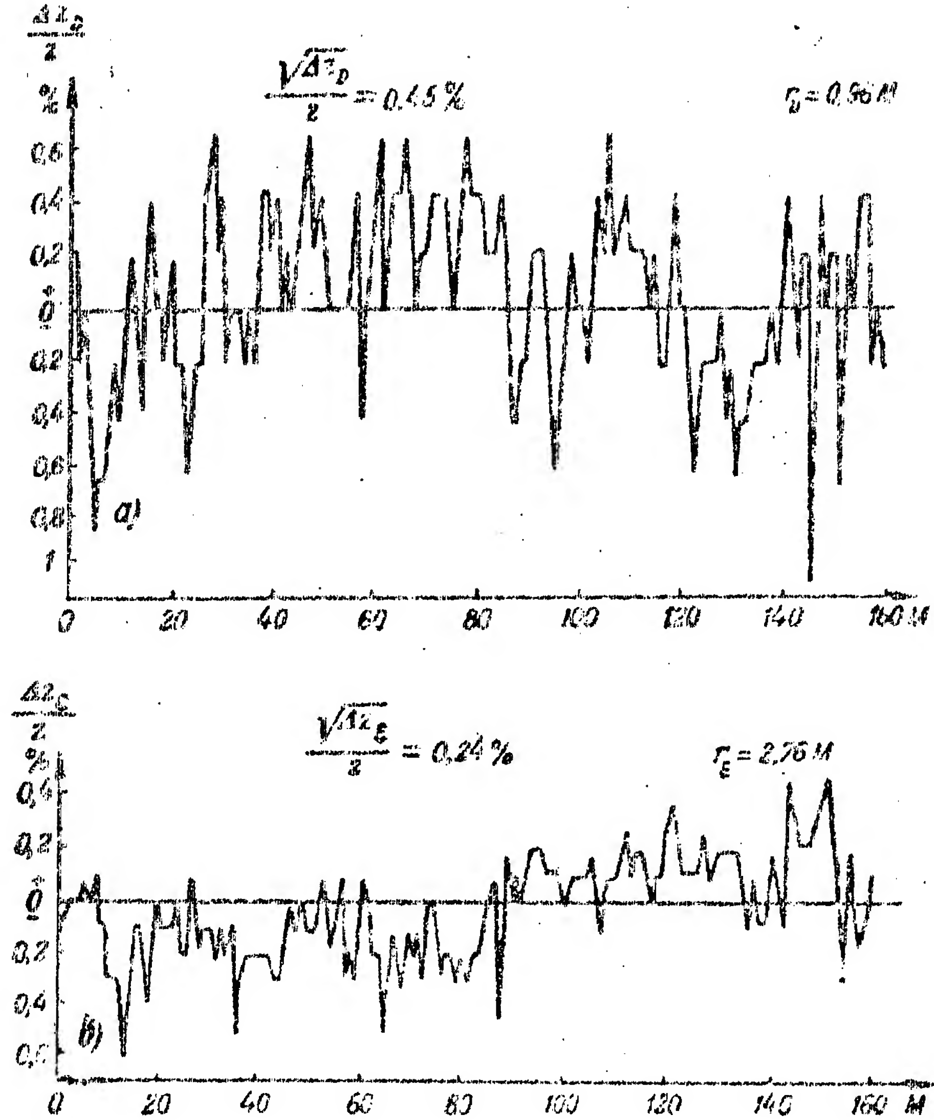


Fig. 5-47a. A) ripple owing to nonuniformity of outer conductor; B) ripple owing to nonuniformity of dielectric.

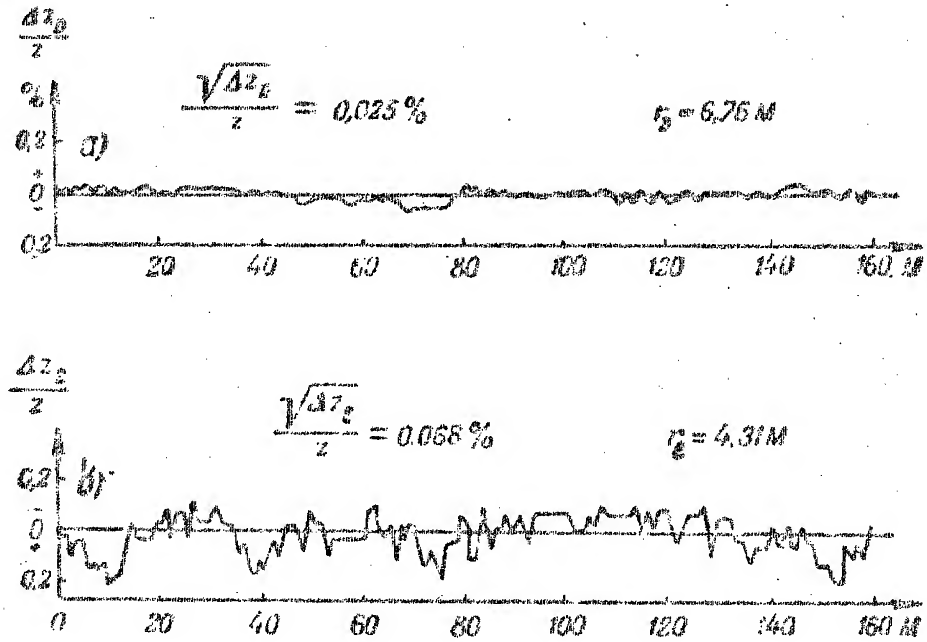


Fig. 5-47b. A) ripple owing to nonuniformity of outer conductor; B) ripple owing to nonuniformity of dielectric.

Variation in the magnitude of the resultant dielectric constant, associated primarily with changing volume of the solid dielectric along the length of the cable, also has a considerable effect on the parameters of the cable; thus in many respects the value of Δz_{ε} is determined by the type and structural form of the insulation (beads, spiral, cord, continuous insulation, etc.). Figures 5-47a and 5-47b give relative values of ripple for two types of cable owing to nonuniformity of the outer conductor, $\Delta z_D/z$ and of the dielectric, $\Delta z_{\varepsilon}/z$

In the first cable, which has the common styroflex type of insulation, the outer conductor is made of copper tape, while in the second cable, having styroflex-shell insulation, the outer conductor is of the tubular type.

It is clear from the figure that the second cable has greater uniformity. Its ripple does not exceed 0.2%, while at the same time the ripple reaches 1% in the first cable.

The mean-square deviation of the wave impedance in the cable is $\sqrt{\Delta z_D/z} = 0.45\%$.

and $\sqrt{\Delta z_e/z} = 0.24\%$, while in the second cable they

are, respectively, 0.025% and 0.068%.

Theoretical and experimental studies of coaxial-cable nonuniformity show that its magnitude and nature may be expressed in terms of the so called "correlation distance" r. This is the distance at which neighboring nonuniformities cease to be independent.

With the aid of the theory of probabilities a formula may be obtained which makes it possible to establish the value of the mean-square variation in the wave impedance of a cable

$$\Delta z_l^2 = \Delta z^2 \frac{2r^2}{l^2}\left(\frac{l}{r} - 1 + e^{-\frac{l}{r}}\right), \qquad (5\text{-}56)$$

where r is the correlation distance (1-5 m on the average);

l is the length of the sections of cable, m;

Δz^2 is the mean square deviation for infinitely small sections at $r = 0$ (the limit case).

From Fig. 5-48, where we give the mean-square variation in wave impedance for various lengths of cable and differing r, it follows that as the length l increases, the mean-square variation z decreases, and the less the correlation distance the greater the value of the variation z. The more frequently points of nonuniformity occur along the cable, the greater the ripple of the cable.

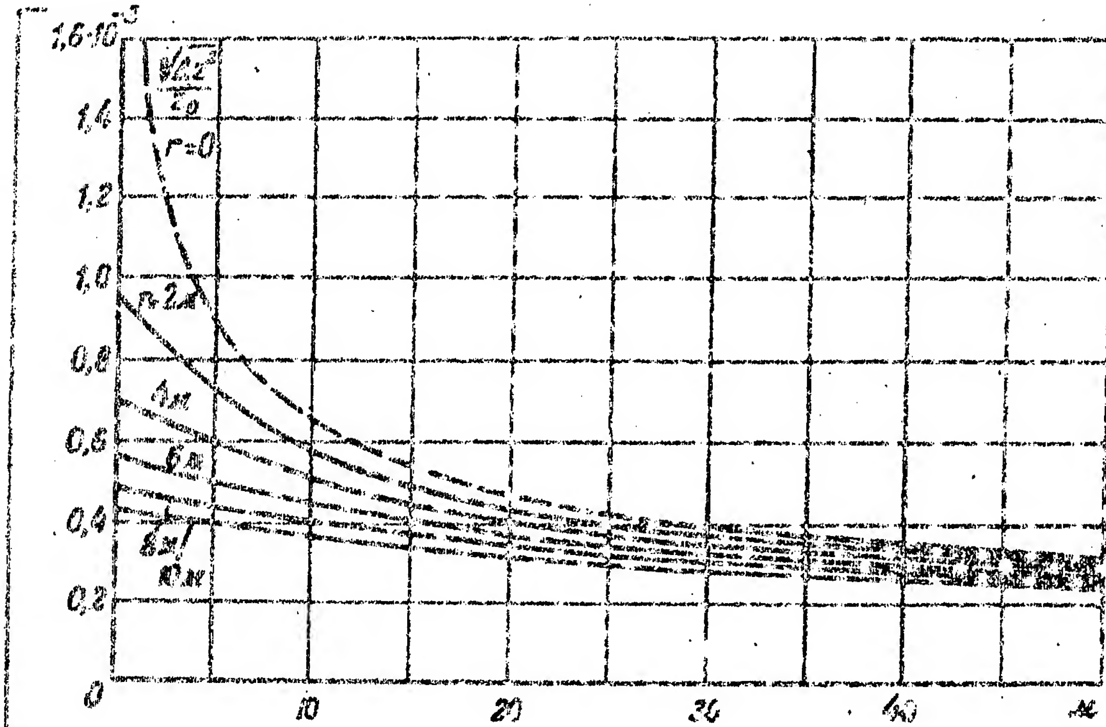


Fig. 5-48. Nonuniformities in a cable for various correlation distances.

The vaule of the forward current may be computed using the following formula:

$$|\Gamma_{\psi}|^2 = \frac{\alpha^2\sqrt{L}}{\sqrt{\beta}\,(1+4\alpha^2 r^2)} \cdot \frac{\Delta z^2}{z^2}, \qquad (5\text{-}57)$$

where α is the phase constant;

β is the attenuation;

L is the length of the transmission path;

r is the correlation distance.

Forward current is increased by a frequency rise, by an increase in the length of the cable link, and by an increase in the nonuniformity.

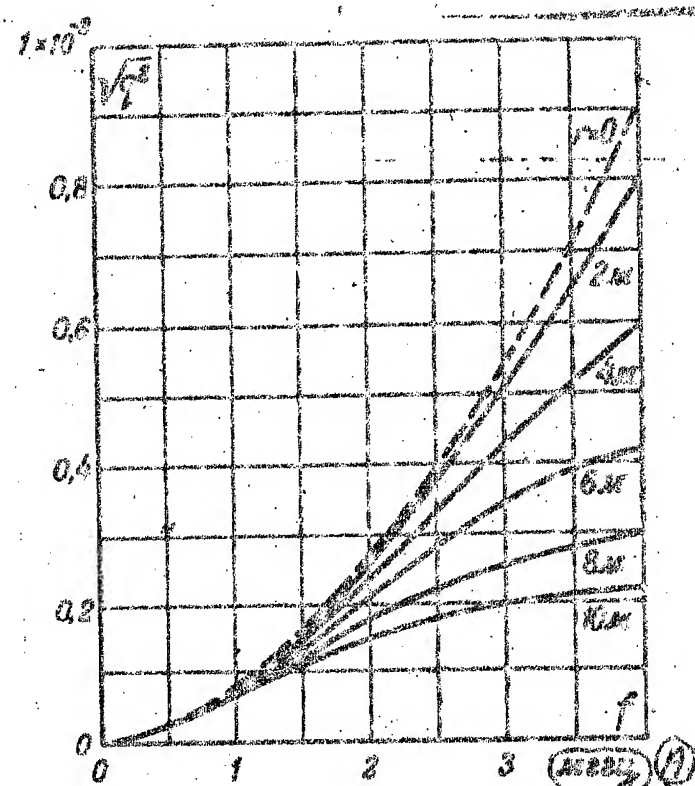


Fig. 5-49. Interference in a cable owing to nonuniformities at various correlation distances.

As β and r increase, the value of q decreases (Fig. 5-49).

Measurement of the input impedance of a cable and a study of its ripple may disclose mechanical damage and defects in the cable. Experience has shown that a sharp break in the regular course of the curve of z_{in} (a peak) is connected with dents in the cable and the point at which they are located. The closer to the beginning of the line that a deformation is located, the

greater its effect on the graph of the input impedance.

It is interesting to note, that dents not exceeding 20 cm in length have very little effect on the characteristic z_{in}, provided that there are not more than 3--4 of them per factory shipping length.

Cables having a large number of small deformations, as well as those with dents of great length, have completely unsatisfactory electrical characteristics.

Repeated comparison tests of cables with beads and with styroflex spiral cord insulation shows that the former are more uniform and that their characteristics are more constant. The results of measurements of nonuniformity for different types of cables are given in Table 5-17.

TABLE 5-17.

Results of measurements of nonuniformity in cables having bead and spiral-cord insulation. A) type of cable; B) insulation; C) maximum values of $\Delta z/z$,% D) average values of $\sqrt{\Delta z^2}/z$, % ; E) factory shipping length, m; F) "Frekventa" beads; G) polyethylene beads; H) styroflex cord; I) "Frekventa" beads; J) styroflex cord.

Тип кабеля	Изоляция	Максимальные значения $\frac{\Delta^2}{z}$, %	Средние значения $\frac{\sqrt{\Delta^2}}{z}$, %	Строительная длина, *м*
2,6/9,4	Шайбы из фреквенты	1,8	0,7	425
2,6/9,4	Шайбы из полиэтилена	1,3—2,7	0,4—0,8	425
2,6/9,4	Стирофлексный кордель	2,9—4,4	1—2	425
5/18	Шайбы из фреквенты	2,3—3,8	—	281
5/18	Стирофлексный кордель	2,9—7,06	—	281

For coaxial-cable communications, it is necessary to consider, in addition to the internal nonuniformities, the nonuniformities at joints as well, which are caused by variations in the characteristics of the factory lengths which are being joined.

In different coaxial cables there are differing relationships between the joint (z_j) and internal (z_{int}) nonuniformities. It has been established experimentally that in new cables the nonuniformities at joints, as a rule, exceed the internal nonuniformities, and the magnitude of z_j/z_{int} amounts to 1--3 in a factory shipping length.

Conversly, in deformed cables, the internal nonuniformities exceed the nonuniformities at joints, and z_j/z_{int} = 0.01--0.1.

In order to make the electrial characteristics

of coaxial trunks more uniform, the factory shipping lengths of cable are specially grouped prior to being laid. They are so grouped that the wave impedance increases from the beginning of the repeater section to its middle and drops from the middle of the section to its end. This is done in such a manner that the deviation of wave impedances between any two adjacent sections of cable does not exceed 0.2%.

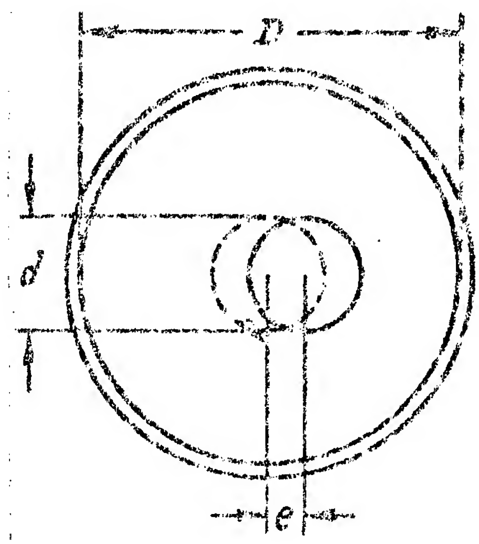


Fig. 5-50. Inner conductor of a coaxial cable located eccentrically.

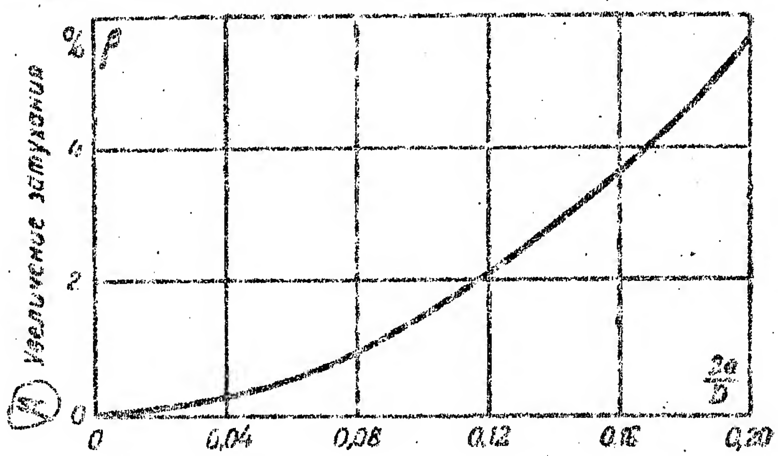


Fig. 5-51. Increase of attenuation with eccentric location of inner conductor. A) increase of attenuation.

The eccentricity e of the inner conductor of a coaxial cable relative to the outer conductor (Fig. 5-50) also has an effect upon the variation in the input impedance, z_{in}, and attenuation, β.

The ripple in a coaxial cable owing to eccentricity is computed by the formula:

$$\frac{\Delta z_e}{z} = \frac{4e}{D^2-d^2} \cdot \frac{1}{\ln \frac{D}{d}}, \qquad (5\text{-}58)$$

Where e is the eccentricity of the inner conductor of the cable. From Fig. 5-51, which gives the percentage increase of the attenuation of a coaxial cable owing to eccentricity, it follows that in practice the eccentricity is small and the increase in attenuation inconsiderable. The magnitude of the nonuniformity in cable lines used for

long-distance communications is regulated by the recommendations of the International Consultative Committee.

According to the standards of the International Consultative Committee, for all types of coaxial cable trunks the attenuation of a nonuniformity must be not less than 5 nepers, i.e.,

$$b_{\text{нонунь}} = \ln \frac{100}{x} = 5 \quad \text{Nepers}$$

This means that the coefficient of reflection, x, may not exceed 0.65%. For very long trunks, the attenuation of a nonuniformity is taken to be 6 nepers, corresponding to x = 0.25%.

The vaules of nonuniformity may be made somewhat more specific with respect to various types of cables. For large coaxial cables the following requirements exist:

A) A joint nonuniformity for any two adjacent factory length sections of cable must not exceed 0.2% at 10^6 cps, that is, if z = 70 ohms, then the permissible value of deviation of the wave impedance, Δz, is 0.14 ohms per shipping length of cable;

B) The fluctuation of wave impedance of the cable over a repeater section for the frequency band used must not exceed plus or minus 4% ($\Delta z = \pm 2.8$ ohms).

For small coaxial cables the following nonunifor-

mity standards have been established:

A) The wave impedance of a shipping length of cable must lie within the limits $z = 75.1 \pm 0.2$ ohms;

B) The mean square difference between the of any coaxial pair and the average value of z for all the shipping lengths sections of cable must not exceed 0.2%.

C) The mean squared differences for z, measured from both ends of the coaxial pairs with respect to the average z of all the shipping lengths of cable must not exceed 0.3%.

In order to meet these standards the following requirements are placed upon the outer conductor and insulation of a small coaxial cable:

A) The permissible variation in thickness of the outer conductor of the cable shall not exceed 0.5 ± 0.05 mm;

B) The thickness of the insulating beads shall not deviate from the nominal value by more than $\pm$ 0.05mm.

For 1.6-mm thick beads, the tolerance is $\pm$ 3.1%.

At present, the investigation of nonuniformities in coaxial cables is carried out mainly by the pulse method, using very sensitive pulse instruments. The instrument permits observation on its screen of the degree of non-uniformity of wave impedance for a cable along its length, and also permits establishment of the site and nature

of a cable defect.

5-13. STANDARDS OF THE INTERNATIONAL CONSULTITIVE COMMITTEE FOR TRUNK-TYPE COAXIAL CABLES.

The International Cousultitive Committee on Telephony in 1946 adopted, in part, the following recommendations for trunk coaxial cables.

1. Type of coaxial pairs.

Diameter of inner conductor 2.6 mm

Inner diameter of outer conductor 9.4 mm.

Thickness of outer conductor 0.25 mm.

Inner and outer conductors to be made of copper.

The surface of the coaxial pair to be covered by a shield in the form of spirals of two steel tapes.

2. The cable should effectively transmit a frequency band ranging from 60 kc to 2,540 kc. The monitoring frequency is 2,852 kc.

3. The attenuation per kilometer at a frequency of 2,500 kc and 15° C should not exceed 0.47 nepers. The distance between repeater points is 9.7 km for a transmission frequency of 2,540 kc.

4. The wave impedance of the cable is 75 ohms at a frequency of 1 Mc.

5. The interference resistance at the transmitting end between two coaxial pairs of the cable shall be not

less than 9.8 nepers over the entire effectively transmitted frequency spectrum. This is defined at zero absolute power level at the input of the circuits (both for the disturbing and for the disturbed circuits).

6. At a frequency of 2 Mc, the average value of wave impedance of all coaxial pairs must lie within the 74.9 to 75.3 ohm range. The mean square difference between the value of wave impedance of a coaxial pair and the average for the remaining pairs shall not exceed 0.2%.

7. The insulation must be subjected to a 2,000 v 50-cycle voltage applied between the inner and outer conductors of the cable. Each factory shipping length of cable must be tested.

8. The resistence of the insulation between the inner and outer conductors most not be less than 5,000 megohm per km after the application of not less than 500 v for 1 min at a temperature not below 10° C.

5-14. TRUNK-CABLE POWER SUPPLY

In modern carrier systems for multiplexing trunk cables the length of a repeater section is only 10--20 km on coaxial circuits and 30--40 km on symmetric circuits. This makes it necessary to install repeater points (UP) at every 10--20 km along the cable.

Technical and economic considerations make it

desirable to have two types of repeater points:

a) Attended repeater points (OUP) having the appropriate operating and technical personnel, located every 70--100 km along the trunk line and

b) Unattended repeater points (NUP), operating automatically without attendent personnel: these points are located 10--20 km apart.

Figure 5-52 shows one of the various systems for locating repeater points, types OUP and NUP, along a cable trunk using combination coaxial cable, type 2.6/9.4.

Here the attended repeater points are located 80 km apart, while the unattended repeater points for the coaxial cable are placed 10 km apart.

A problem which is extremely important for cable links is the question of the power supply for the unattended repeater points (NUP).

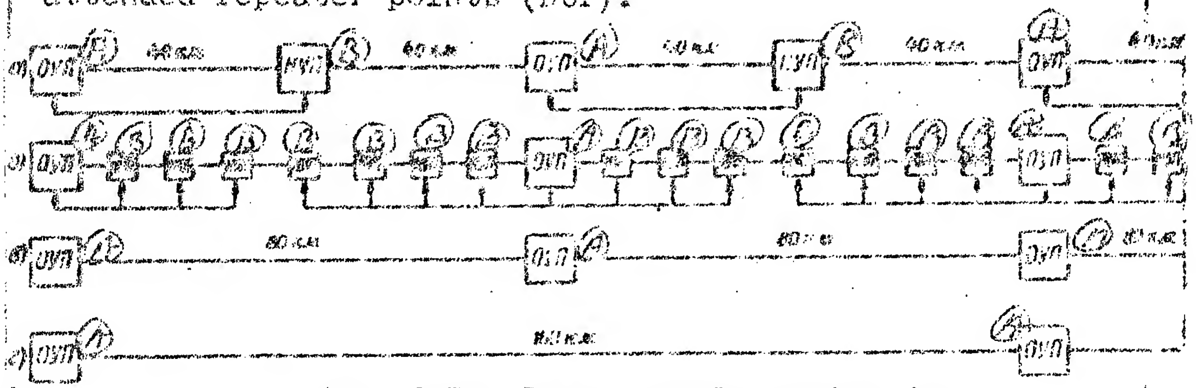


Fig. 5-52. Power supply system for a trunk cable. a) carrier communication over noncoil loaded circuits (12-channel

system); b) Telephone and television transmission over coaxial pair, type 2.6/9.4; c) radio broadcasting (lightly coil-loaded system, L_S = 12 mh, S = 1.7 km); d) low frequency communication (average coil-loaded system, L_S = 100/70 mh; S = 1.7 km). A) attended repeater point; B) unattended repeater point.

There are three possible types of NUP power supply:

1) Equiping the NUP with completely automatic power plants using diesel, turbine, or wind-powered installations;

2) Supplying the NUP from the OUP's over a seperate power cable laid parallel to the communications cable along the entire trunk;

3) A system of supply in which NUP's are supplied from the nearest OUP over current carrying conductors in the same communications link, the so called "remote power supply system."

In comparing the NUP power-supply systems discribed, it should be noted that the first requires complete automation of all the NUP power installations, which is difficult to achieve at present, while the second is undesireable from an economic point of view.

In modern installations for combination coaxial-cable links, the remote power-supply system has come into wide use for the unattended repeater points.

As a rule, 500--1,000-v electrical energy is sent along the conductors of the coaxial cable from an OUP to the neighboring NUP's.

Each OUP serves an appropriate number of NUP's in both directions. From Fig. 5-52 it may be seen that from one OUP power is supplied to 7 NUP's: 3 in one direction and 4 in the other.

As a rule, 50--60-cps AC is used to supply coaxial trunks. In this case, the advantages of AC over DC are as follows:

1) Simplicity in transforming the voltage at the NUP;

2) Less danger of cable corrosion (when a ground-return system is used);

3) Uncomplicated regulation of voltage at the NUP.

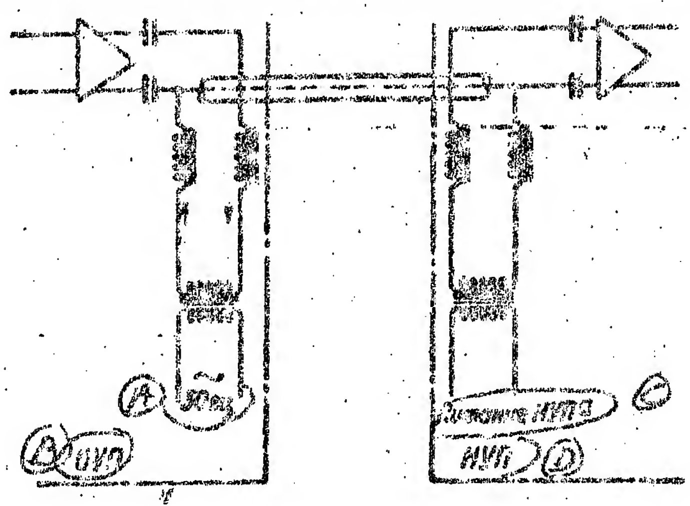


Fig. 5-53. Electrical power supply over the outer and inner conductors of a coaxial cable. A) 50 cps; B) OUP; C) NUP supply; D) NUP.

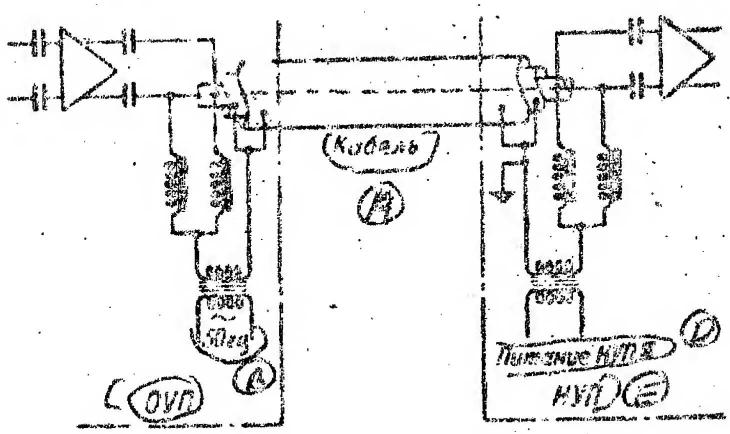


Fig. 5-54. Electrical power-supply along the circuit: forward direction--inner and outer conductors of the cable; reverse direction--lead sheath and armoring of the cable. A) cable; B) 50 cps; (cont'd)

C) OUP; D) NUP supply; E) NUP.

The drawbacks to AC supply are the appearance of some interference on the communications circuit and the necessity of providing power-supply filters at the NUP.

However, the advantages of AC-power supply overshadow the disadvantages, and it is chosen in preference to DC.

There are four ways of setting up remote power supply over a coaxial cable.

I. The electrical energy is transmitted over the inner and outer conductors of the coaxial pair (Fig. 5-53).

In this case power is supplied over the same coaxial circuit which carries communications signals, and it is possible that considerable interference will arise when heavy currents are used. Also, in this case, there will be considerable energy losses in the conductors, since the circuit has a rather high resistance.

II. Power can also be supplied over the circuit which is formed by using the inner and outer conductors of a coaxial cable connected in parallel as the outgoing conductor, and the lead sheath and armoring as the return conductor (Fig 5-54).

The defects in method II for supplying electric power are the large resistance of the return conductor, the dangerous overvoltages which may occur if the armoring at the joints is poorly soldered or if rounding is not satisfactory, and interference of the power current with the communications signals.

III. Electrical power is transmitted to the NUP by using two parallel-connected inner conductors as the forward conductor, and two parallel-connected outer conductors of coaxial cables for the return conductor (Fig. 5-55).

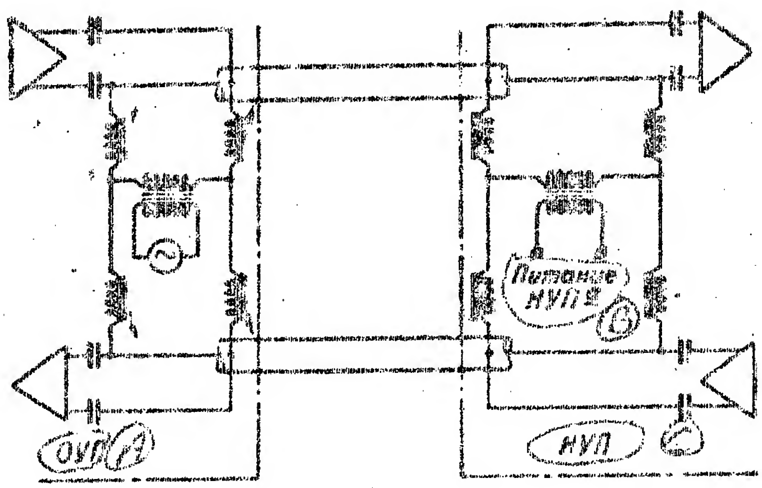


Fig. 5-55. Electrical power supply using the circuit: forward direction -- two parallel-connected inner conductors; return direction -- two parallel-connected outer conductors.

Variant III is basically free from the faults possessed by methods I and II for supplying NUP's; here, however, the supply circuit utilizes two coaxial pairs, and if one of them breaks down, communications over both pairs will be interrupted.

With the four-wire system in use at present, in which the forward and reverse transmissions take place over separate coaxial pairs, this is not a very grave fault, since if one pair breaks down, neither will be used anyway.

IV. Electrical power is supplied to the NUP's over circuits in which two inner conductors of coaxial cables are connected in parallel to form the forward conductor, while the circuit is completed through a ground return (Fig. 5-56). This method is similar to the one just described, except that in this case the resistance of the power-supply circuit is considerably less, and nearly 1.5 times more power may be transmitted over the cable.

It should be kept in mind that in order to insure power-supply stability it is necessary to carry out the installation of the grounding circuits for the power supply at the repeater points with great care.

When the power-supply method described are compared from the technical and economic points of view, variant IV appears to be the best way for supplying unattended repeater points.

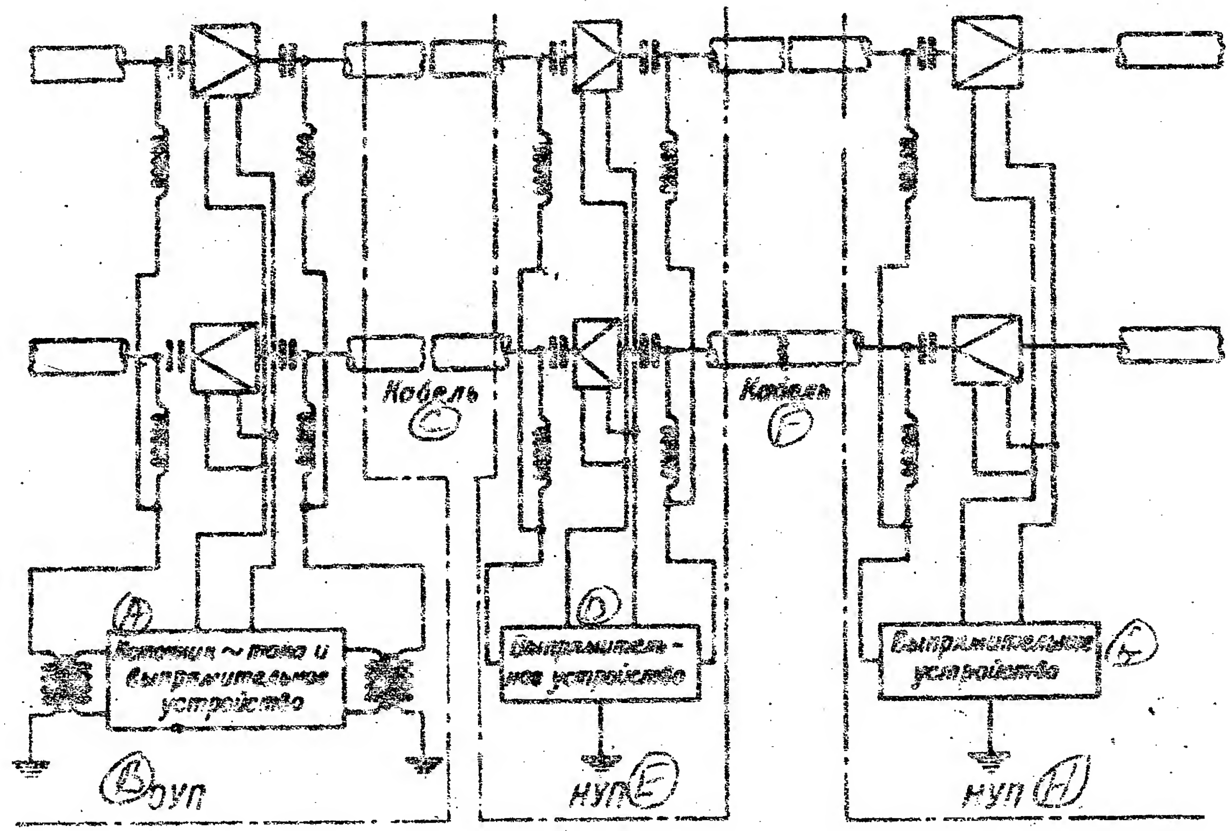


Fig. 5-56. Electrical power supply using the circuit: forward direction — two inner conductors connected in parallel; return direction — ground. A) AC source and rectifying equipment; B) OUP; C) cable; D) rectifying equipment; E) NUP; F) cable; G) rectifying equipment; H) NUP.

For an average coaxial cable, type 2.6/9.4, with the modern degree of carrier multiplexing, it is necessary to supply the equipment of a single NUP with 1200-1500 watts of power at 800 v.

Unattended repeater points of symmetric carrier circuits are also supplied remotely. As a rule, 220-v energy is transmitted to them over the same symmetric circuits as are used for the communications signals.

On other trunks, the following methods are currently used to supply coaxial-cable NUP's.

For small type 1.83/6.7 cable having four coaxial pairs, the NUP's are supplied over a circuit which uses the four inner conductors of the coaxial pairs for the forward conductor, while the return conductor is the ground. 800 watts of power at 280-350 v are transmitted to each NUP. 60-cycle AC energy is transmitted. Each OUP supplies 4-5 NUP's in each direction.

On this type of trunk (1.83/6.7), the NUP's had previously been supplied over the inner conductors of the coaxial pairs without using the ground as the return conductor.

On trunks using the large type 5/18 coaxial cable, the NUP's are supplied in normal operation from the local mains. In case of a breakdown in the local power source,

the NUP's are supplied remotely over the coaxial cable.

In order to supply the equipment of a single NUP, 2000 watts of single-phase 550-v AC power is required. 750 watts are required for heating and ventilation of an NUP.

5-15. COAXIAL SUBMARINE CABLES

The most important requirement associated with submarine cables is that they be able to transmit communications over a long span, using no intermediate repeaters, or a minimum number of them.

It is clear that the use of underwater intermediate repeater stations involves several difficulties when cable trunks are constructed and operated.

Contemporary submarine cables must be suitable for both low frequency and multichannel carrier communications over long distances.

The specific feature of submarine communications cables that is due to the conditions under which the cables are laid and operated in the water, is the high moisture resistance of the insulation and the outer covering of the cable. The stability of the electrical parameters of the cable must remain high over extended periods of operation in the water. The mechanical strength of the cable must be computed on the basis of the various water currents and

pressures at different depths.

A coastal cable must have a specially reinforced armoring. It must resist the action of coastal tides and withstand possible shocks from coastal rocks, anchors, hooks, etc.

The development of submarine cables in recent years has taken the direction of increasing the communications span, and widening the transmitted frequency band, since the first under-water communications were telegraphic, and it has only been since the twenties that cables have been laid which are suitable for long distance telephony.

The setting up of telephone communications under the Atlantic and Pacific oceans is as yet an unsolved problem. An artificial increase in the inductance of the cable circuits, which increases the communications span over a submarine cable 2-4 times, permits only telegraphic transatlantic communications.

The basic drawback to submarine cables is their unsuitability for carrier multiplexing.

Modern requirements for setting up high-frequency communications under large expanses of water are met most fully by the coaxial cable. However, it should be noted that submarine cable trunks are multiplexed over a frequency band not exceeding 60-100 kc, and symmetric cables

could be used for this purpose with no less success. Coaxial cables, however, consume nearly 1.5 times less material than do symmetric cables, to say nothing of the possible increase in the usable frequency range.

The first marine coaxial cable, laid in 1920-1921, had an inner copper conductor, loaded by means of ferromagnetic tape. This increased the span of communications over the cable, but at the same time limited the transmitted frequency spectrum to the voice-frequency range (up to 3000 cps).

Later, as requirements rose in a number of links, artificial increases in inductance were discarded, and standard high-frequency coaxial cables began to be used.

In view of the isolation of underwater cables, and their resistance to external noise, it is permissible to increase the power of the transmitting station considerably for underwater coaxial trunks, and also to use a lower level at the receiving station. As a result, it is possible to increase the attenuation over the span of an underwater cable trunk up to 10 nepers, although this value, as a rule, does not exceed 6 nepers for underground cable trunks.

This makes it possible to carry out high-frequency communications over cables having repeater points 150-250 km apart.

The principle structural peculiarities of submarine coaxial cables, in comparison to underground cables, are:

1. The considerably greater dimensions of the cable; where the diameter of the outer conductor of an underground coaxial cable does not exceed 18 mm, in the submarine cable it may reach 30-40 mm;

2. The current carrying conductors of the cable are flexible; the inner conductor consists of a large-cross section wire and a layer of thin circular wires or tape.

The outer conductor is made of flat tape or of circular wires.

Figure 5-57 shows typical cable structures. The structural dimensions and electrical data for the cables are given by Table 5-18.

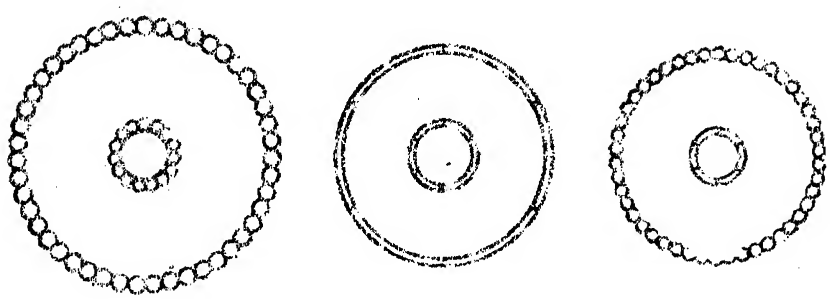


Fig. 5-57. Construction of coaxial submarine cables.

A Номер кабеля	B Внутренний провод			J Диэлектрик		P Внешний провод		V Затухание при $f = 40$ кгц, мнеп/км	W Волновое сопротивление, ом
	C Сплошной провод диаметром, мм	D Повив лент или проволок		K Материал	L Радиальная толщина, мм	Q Количество	R Диаметр или толщина, мм		
		E Количество	F Диаметр или толщина, мм						
I	2,51	10 проволок G	1,05	M Резина	6,35	48 проволок S	0,940	63,4	54,0
II	3,5	6 лент H	0,36	N Парагутта	5,70	6 лент T	0,482	56,0	51,4
III	3,1	5 лент I	0,40	O К-гутта	4,35	34 проволоки U	0,890	92,0	44,0

Table 5-18. Structural and Electrical Data for Marine Coaxial Cables. A) Cable No.; B) inner conductor; C) diameter of solid conductor, mm; D) layer of strips or wires; E) quantity; F) diameter or thickness, mm; G) wires; H) strips; I) strips; J) dielectric; K) material; L) radial thickness, mm; M) rubber; N) paragutta; O) K-gutta; P) outer conductor; Q) quantity; R) diameter or thickness, mm; S) wires; T) strips; U) wires; V) attenuation at f = 40 kc, mnepers/km; W) wave impedance, ohms.

3. The insulation is made in the form of a continuous shell; for the first type of cable structure, intended for telegraph communications, guttapercha is chiefly used. It has the advantages of good electrical insulating properties, excellent moisture resistance, and good elasticity. Its defect is the occurrence of considerable dielectric losses in the high-frequency spectrum, making it difficult to use guttapercha for high-frequency telephone communications.

In addition the properties of guttapercha deteriorate sharply under the action of atmospheric factors, which substantially complicates the storage and transportation of the cables.

Among the dielectrics which have somewhat better properties (lower tan δ), than guttapercha, and somewhat better for use in marine coaxial cables, are:

a) paragutta — a mixture of guttapercha, purified wax, and deproteinized rubber;

b) K-gutta — a mixture of guttapercha and petroleum jelly;

c) special rubber-type composition.

In recent designs of submarine coaxial cables, polyethylene has been used for the dielectric.

4. There is a specially reinforced protective ar-

moring, protecting the core of the cable, and giving it the tensile strength required in laying the cable and in possible retrievals for repairs.

Cables are classified as deep-sea and coastal on the basis of their armoring.

Deep-sea cables are armored with round steel wires, 2-5 mm in diameter. High-breaking-strength steel is used.

As a rule, the armoring of coastal cables consists of two layers of round steel wire.

The jute covering of the cable is treated in tannin, so that the jute will not contract in the sea water.

Below we give some quite characteristic designs of submarine coaxial cables and their basic electrical characteristics.

1. Type 7/25 Cable (Fig. 5-58)

The flexible inner conductor of the cable consists of a central copper core, 4 mm in diameter, and a layer of copper conductors, 1.5 mm in diameter. The diameter of the inner conductor is 7 mm. The polyethylene insulation is 9 mm thick. The outer conductor is made of copper wires, of flat cross section, 6 by 0.6 mm. On top of this there is a reinforcing tape of annealed copper. Next comes the protective shell of polyvinyl chloride. Finally there is the

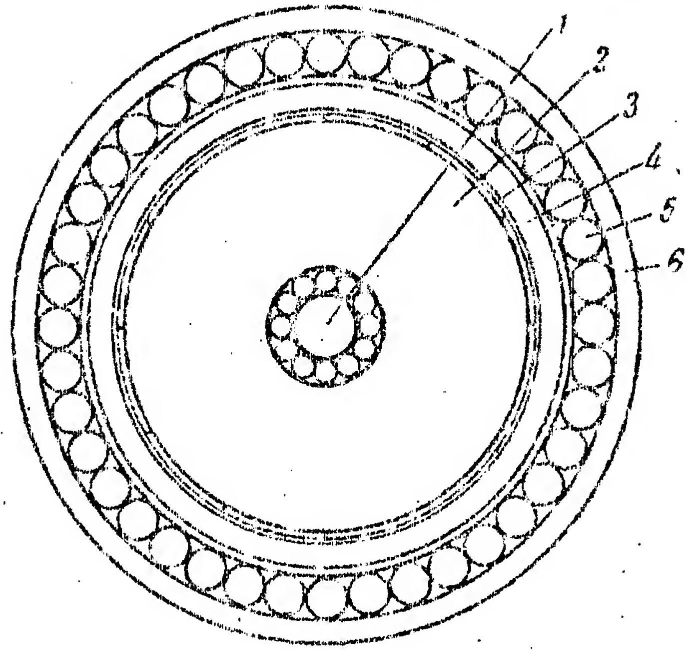


Fig. 5-58. Submarine coaxial cable, type 7/25. 1) Inner conductor (copper); 2) insulation (polyethylene); 3) outer conductor (copper); 4) protective shell (polyvinylchloride); 5) armor (steel); 6) jute.

jute padding and the steel-wire armoring — the wires being 4-5 mm in diameter.

The coastal section of the cable uses armoring of 6 mm diameter wire.

Above the armoring of the cable there is yarn and a chalk solution.

The cable is shipped in lengths of up to 10 kilometers.

The frequency-dependence of the cable attenuation is shown in Fig. 5-59.

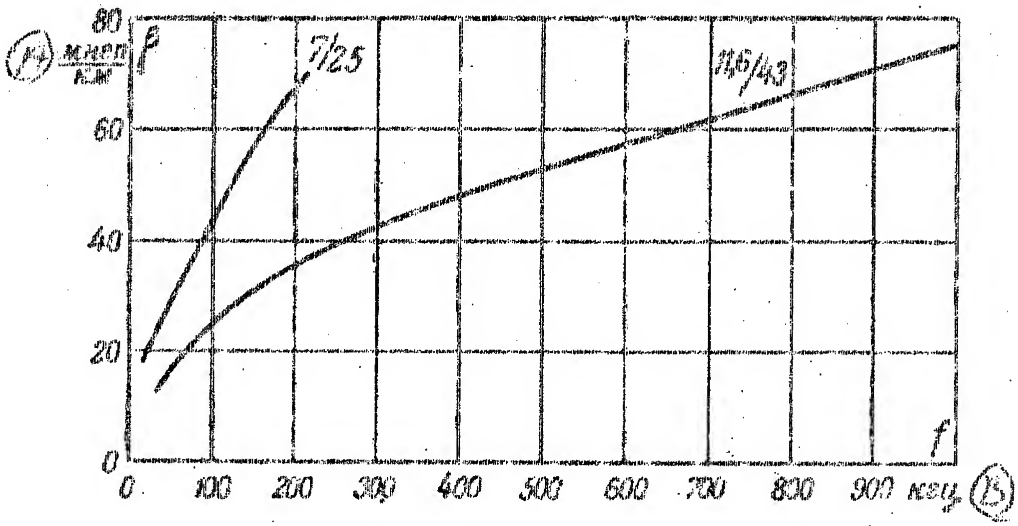


Fig. 5-59. Attenuation per kilometer of submarine coaxial cables. A) mnepers/km; B) kc.

The cable is multiplexed in a band extending to 60,000 cps.

The 300 to 3000 cps frequency band is set aside for voice-frequency telephony or for 12-18 voice-frequency telegraph links.

The 32,000 to 60,000 cps band is used for six telephone carriers in what amounts to a four-wire system (half of the band is used for transmission in one direction, and the other half for transmission in the opposite direction).

The communications system is single cable. The distance between repeater points is about 200 km.

Similar cable designs are known, which are multiplexed up to 108 kc in a two-cable system.

In this frequency range, using two cables, 24 telephone carriers are available.

Telephone transmission is also carried out over carrier channels; in this case each telephone channel is replaced by 18 telegraph channels.

II. Type 11.6/43 Cable (Fig. 5-60)

This cable has three conductors which are mutually concentric and form two coaxial circuits.

The first coaxial circuit consists of the inner and outer conductors, and is used for multichannel telephone and telegraph communications. To set up the second circuit, the inner conductor of the first circuit and the central conductor of the coaxial cable are used. This circuit is used for monitoring, for technical operating measurements, for signalling, and for service communications along the trunks.

The structure of the cable is as follows:

1) The central conductor consists of a copper conductor, 1.7 mm in diameter, and a layer of 10 copper wires,

0.76 mm in diameter each. The conductor is overlaid with a layer of polyethylene having an outside diameter of 11.22 mm.

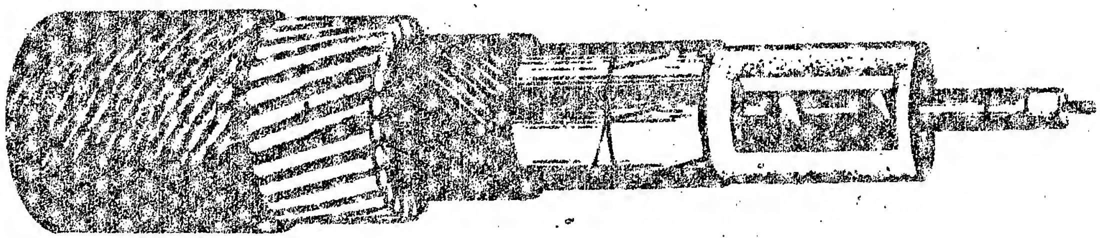


Fig. 5-60. Over-all view of the type 11.6/43 marine coaxial cable.

2) The inner conductor consists of six copper wires, laid spirally, with a single covering tape. The thickness of the inner conductor is 0.19 mm.

The outside diameter of the inner conductor is 11.6 mm.

3) A combination of insulations is used.

A polyethylene cord, 5.5 mm in diameter, is wound spirally about the inner conductor, with a pitch of 25.4 mm. Then a cylindrical polyethylene tube having an outside diameter of 43 mm is placed over the cord.

4) The outer conductor consists of six copper tapes, 0.38 mm thick, wound spirally, and two open spiral coverings of wide copper tape, 0.1 mm thick.

5) The protective armor consists of a tarred jute cushion, armor of 23 wires, a layer of compound, and an outer jute covering.

The cable weighs 11.8 tons (metric)/km.

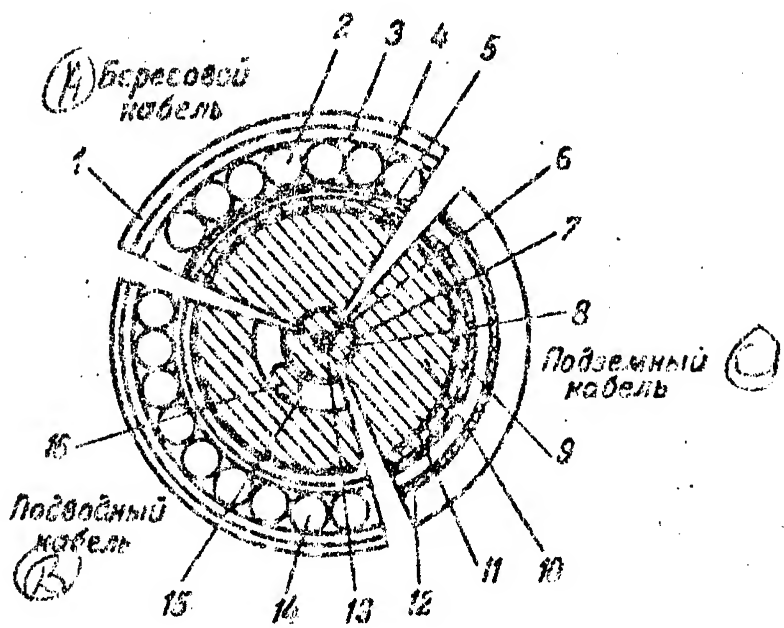


Fig. 5-61. Structure of type 11.6/43 submarine cable. 1) Two layers of jute; 2) wire armoring; 3) compound; 4) two layers of jute; 5) outer conductor; 6) central conductor; 7) inner conductor; 8,9,10) polyethylene insulation; 11) lead sheath; 12) tape-type armoring; 13) polyethylene insulation; 14) wire armoring; 15) air gap; 16) polyethylene cord. A) Coastal cable; B) submarine cable; C) underground cable.

Coastal cable, laid to a distance of approximately 1 km from the sea, and also along the coast up to the terminal amplifying station, has continuous polyethylene in-

sulation and reinforced armoring.

In order to equalize the wave impedances of the coastal and submarine cables, the inside diameter of the coastal section is increased by approximately 20%.

The underground portion of the cable differs from the coastal portion only in having a zinc sheathing, and tape- rather than circular-wire-type armoring.

The wave impedance of the submarine cable is 62 ohms. The attenuation of the cable is shown in Fig. 5-59.

The construction of the submarine, coastal, and underground sections of type 11.6/43 cable is shown in Fig. 5-61.

III. The Transatlantic Cable

The transatlantic cable trunk is intended for carrier telephone-telegraph communication between North America and Europe. At present this trunk is in the design stage.

Of interest are some of the technical decisions made in developing the trunk.

First variant. A coaxial cable with solid polyethylene insulation.

The cable is multiplexed by 24-channel carrier equipment for telephony, with a range of up to 100-120 kc. The distance between repeater substations is 100 km.

The amplifiers are located underwater and use long-life electron tubes, permitting uninterrupted operation of the trunk for several decades. Power is supplied to the trunk over the same coaxial cable from the terminal point.

Second variant. The distinctive feature here is that the amplifying equipment is located in the cable itself.

The amplifying elements, tubes, circuits, and so forth are located along the cable in flexible cylindrical joints.

The joint is made of a series of steel rings, located under the cable armor. Each ring is 38 mm in diameter, and 19 mm wide. Above these rings there is another series of thinner steel rings, covering the joints between the rings of the first group. Assembled in this way, the rings form an articulated cylinder about 2 m long. The cylinder is carefully sealed to prevent moisture from penetrating.

All in all, the cable is flexible enough so that it may be mounted on drums and laid from them.

It is proposed that polyethylene or paragutta be used for insulation.

The central conductor of the cable weighs 125 kg/km; the dielectric weighs 90 kg/km, the return conductor weighs 145 kg/km.

The usable frequency range extends to 48,000 cps.

The distance between amplifiers is 68 km. It is proposed that 47 amplifier joints be installed along the 3000 km run of the cable.

Power is to be supplied to the submarine cable over the inner conductor of the cable, at 2000 v; the voltage is to be supplied by batteries at the cable terminals.

CHAPTER SIX

INTERACTION AND NOISE RESISTANCE OF SYMMETRIC CABLE LINKS

6-1. THE NATURE OF INTERACTION IN COMMUNICATIONS CABLES

The process of propagation of electromagnetic energy along conductors was considered in Chapter 2. However, in designing and manufacturing high-quality communications cables, it is also necessary to study the phenomenon of the transfer of energy from one circuit to another, and the ability of the cables to resist interference.

The ability of cables to withstand interference is a very important factor in providing reliable communications service, and is especially significant in long-distance high-frequency telephony and telegraphy. In this case, the quality and distance of a link depend not so much upon the attenuation in the cable circuit itself, as upon the interference between adjacent circuits.

It was shown above that at the permissible cable line attenuation of 3.3 nepers, only 1/735 of the energy sent into the line arrives at the receiver. The major part of the energy (734/735) is dissipated in the cable

itself, chiefly in heat losses and dielectric polarization; in addition, some energy may be transferred into adjacent circuits, appearing as noise currents.

The transfer of energy from one circuit to another depends upon the electromagnetic interaction between them.

When an alternating current is passed through the first (influencing) circuit, there is created around it an electromagnetic field which changes in direction.

As a result of this field, an emf is induced in the second circuit (the affected circuit) located alongside; the emf gives rise to a current that appears in this circuit as interference.

As a result, a transmission sent through the first circuit will be audible to some degree or another over the second circuit.

The electrical and magnetic interaction between the circuits are characterized, respectively, by the capacitive coupling coefficient, C_{12}, and the inductive coupling coefficient, M_{12}.

In general form, these coefficients (Fig. 6-1) are complex quantities, and are expressed as:

$$C_{12} = g + j\omega c,$$

$$M_{12} = r + j\omega m,$$

where g is the active component of the capacitive-coupling coefficient, and is called the "dielectric coupling."

r is the active component of the inductive-coupling coefficient, and is called the "resistive coupling."

c is the capacitive coupling.

m is the inductive coupling.

The quantities g, r, c, m are called the primary interaction parameters.

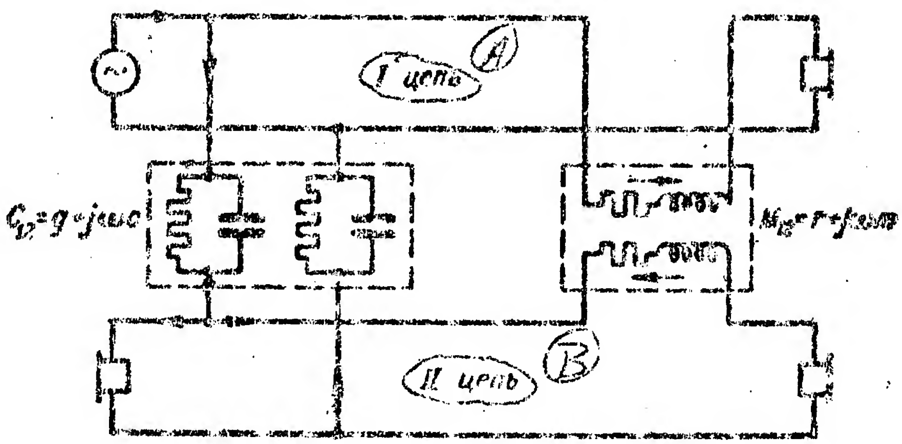


Fig. 6-1. The coefficients of capacitive (C_{12}) and inductive (M_{12}) coupling for a cable. A) First circuit; B) second circuit.

Just as the laws of the propagation of energy along conductors are determined by the four primary transfer parameters R, L, C, G, the process of energy transfer between circuits depends upon the four primary magnitudes r, g, c, and m (Fig. 6-2).

From Fig. 6-2 it can be seen that there is a capacitive interaction circuit $g + j\omega c$ located at the path along which current passes from one circuit to the other, while an inductive interaction circuit $r + j\omega m$ acts in parallel.

For the sake of comparison, the transmission equivalent circuit is shown. It differs in that the resistance-inductance is connected in series, while the conductance-capacitance circuit is connected in parallel.

A secondary parameter of interaction is the quantity B (the cross-talk attenuation), which characterizes the attenuation of the interaction currents upon passing from the first circuit to the second.

The parameter B is similar to the circuit attenuation $b = \beta l$, upon which depends the degree of attenuation of the energy sent through a cable.

In principle, the transfer parameters R, L, C, G and the interaction parameters r, g, c, m differ in that the first have finite values determined by the structure of the cable and the frequency of the current transmitted, while the second can be reduced nearly to zero by proper mutual location of the affected and affecting circuits.

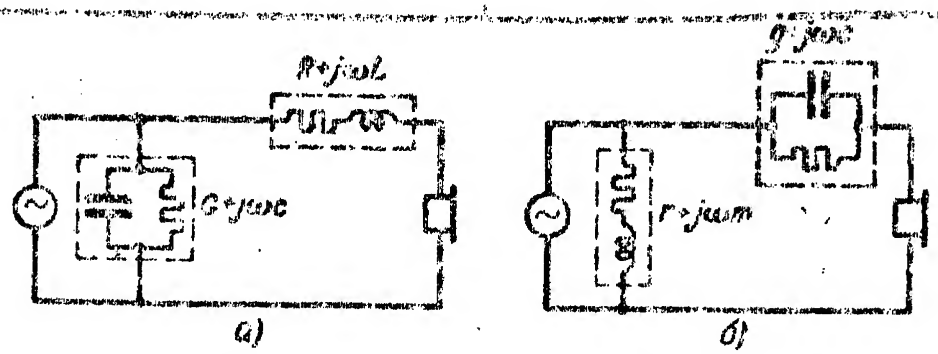


Fig. 6-2. a) Equivalent transmission circuit; b) Equivalent interaction circuit.

Cable designers as a rule try as much as possible to decrease the attenuation of the cable itself, βl, and to increase the cross-talk attenuation B.

The cross-talk attenuation is expressed in terms of the logarithm of the ratio of the power P_1 of the generation feeding the disturbing circuit to the power P_2 of the noise in the disturbed circuit; it is measured in nepers:

$$B = \frac{1}{2} \ln \frac{P_1}{P_2}. \tag{6-1}$$

It may also be calculated in terms of the ratio of the voltages or currents in the disturbing and disturbed circuits:

$$B = \ln \frac{U_1}{U_2} = \ln \frac{I_1}{I_2}. \tag{6-2}$$

In considering the interaction of communication circuits, two types of energy transfer are distinguished: near-end and far-end.

The effect occurring at the end of the circuit where the generator of the first circuit is located, is called the near-end energy transfer, P_{20}. The effect at the opposite end of the second circuit is called the energy transfer at the far end, P_{21}.

Accordingly, the cross-talk attenuation based on power will be (Fig. 6-3):

At the near end

$$B_0 = \frac{1}{2} \ln \frac{P_{10}}{P_{20}}, \tag{6-3}$$

At the far end

$$B_l = \frac{1}{2} \ln \frac{P_{10}}{P_{2l}}. \tag{6-4}$$

In addition to the quantities B_0 and B_l, the parameter B_{12} (circuit noise-resistance) is widely used in communications technology; it is the difference between the levels of the useful signal and the noise (transfer currents), arising at the far end of the cable line (Fig. 6-3)

$$B_{12} = \frac{1}{2} \ln \frac{P_{1l}}{P_{2l}}, \tag{6-5}$$

It can be shown that the noise resistance B_{12} is numerically equal to the difference between the cross-talk attenuation of the cable and the cable attenuation $b = \beta l$

$$B_{12} = B_l - \beta l. \tag{6-6}$$

Actually, since

$$\beta l = \frac{1}{2} \ln \frac{P_{1 0}}{P_{1 l}},$$

then

$$B_l - \beta l = \frac{1}{2} \ln \frac{P_{1 0}}{P_{2 l}} - \frac{1}{2} \ln \frac{P_{1 0}}{P_{1 l}} = \frac{1}{2} \ln \frac{P_{1 l}}{P_{2 l}}.$$

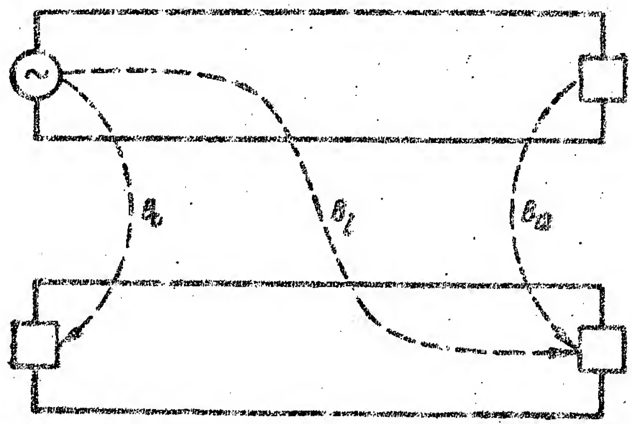
The coupling coefficients are chiefly used in finding the effect of interaction in short lengths of cable on the order of shipping length (a hundred meters).

It is extremely difficult to determine the coupling coefficients in long cable lines, and in many cases it is generally impossible, and interaction is evaluated by using the cross-talk attenuation.

The coupling coefficient, and consequently, the cross-talk attenuation, and accordingly, the degree of interaction of the circuits, depend upon the location of the conductors of the disturbing and disturbed circuit with respect to each other, upon the communications system (single-wire or double-wire), upon the type of lay-up (quadded, paired, twin-paired), upon the uniformity of structure along the cable, and upon the cross-section and quality of the materials used. In addition, interference effects depend upon the length of the cable circuit and the frequency of the signals transmitted.

At present, single-wire circuits are used only in the field of DC telegraphy.

First circuit - the influencing one



Second circuit - the affected one

Fig. 6-3

Interaction between communication circuits.

For links using AC, two- and four-wire circuits are used exclusively, for the reason that single-wire circuits afford no protection against mutual interference.

6-2. EQUATIONS FOR EFFECTS IN SHORT LENGTHS OF CABLE

The analysis given below is correct for short lengths of cable, of the order of shipping length, where the variation of current and voltage along the conductors can be neglected; in this case, only the effect of the first circuit upon the second need be considered; the effect of the second circuit upon the first can be disregarded upon the assumption that it is negligible.

Both the disturbing and disturbed circuits have matched loads at the ends, equal to the wave impedance of the cable, Z.

Let us consider separately the effect due to the electrical field, expressed by the capacitive-coupling coefficient C_{12}, and the effect due to the magnetic field, expressed by the inductive-coupling M_{12}. In the disturbed circuit, the capacitive coupling causes a capacitive-noise current (Fig. 6-4), while the inductive coupling causes an an inductive-noise current.

The capacitive-noise current along the path from the first circuit to the second encounters an impedance

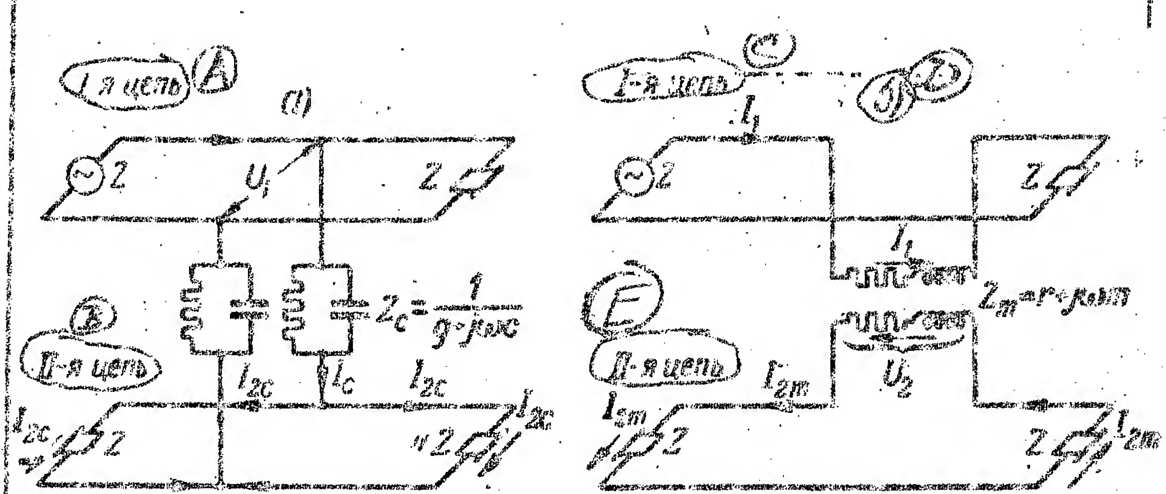


Fig. 6-4. Diagrams for the equations of short-cable effects. a) Capacitive effect; b) inductive effect; A) first circuit; B) second circuit; C) first circuit; D) b; E) second circuit.

consisting of the capacitive-coupling impedance ($Z_c = 1/(g + j\omega c)$) and one-half the wave impedance of the cable ($Z/2$).

The quantity $Z/2$ occurs because the current entering the second circuit splits in parallel in two

directions: toward the near end and toward the far end of the cable

$$\dot{I}_c = \frac{\dot{U}_1}{Z_c + \frac{Z}{2}} = \frac{\dot{I}_1 Z}{\frac{1}{g + j\omega c} + \frac{Z}{2}},$$

where $\mathring{U}_1$ and $\mathring{I}_1$ are the voltage and current in the first circuit.

Considering that $Z_c \gg Z/2$, we may write

$$\dot{I}_c = (g + j\omega c)\, \dot{I}_1 Z.$$

The current in the second circuit, I_{2c}, which is directed toward the near end or toward the far end is equal to one-half the current $\mathring{I}_c$

$$\dot{I}_{2c} = \frac{\dot{I}_c}{2} = \frac{1}{2}(g + j\omega c)\, \dot{I}_1 Z. \qquad (6\text{-}7)$$

The ratio of the current $\dot{I}_1$ in the first circuit to the capacitive-noise current I_{2c} in the second circuit is expressed in the form

$$\frac{\dot{I}_1}{\dot{I}_{2c}} = \frac{2}{(g + j\omega c) Z}.$$

The voltage arising in the second circuit owing to the inductive coupling, $\dot{U}_{2m}$, is determined by the equation

$$\dot{U}_{2m} = \dot{I}_1 Z_m = \dot{I}_1 (r + j\omega m),$$

where $\dot{I}_1$ is the current in the first circuit;

Z_m is the inductive-coupling-impedance.

The inductive-noise current arising as a result of this voltage in the second circuit, $\dot{I}_{2m}$, is closed in series with the near and far ends of the circuit, and thus sees an impedance equal to 2Z

$$\dot{I}_{2m} = \frac{\dot{U}_{2m}}{2Z} = \frac{1}{2}(r + j\omega m)\frac{\dot{I}_1}{Z}. \qquad (6\text{-}8)$$

Then the ratio of the current in the first circuit to the induced-noise current in the second circuit will be

$$\frac{\dot{I}_1}{\dot{I}_{2m}} = \frac{2Z}{r + j\omega m}.$$

As has been shown above, the relation of the currents in the first and second circuits is characterized by the cross-talk attenuation B.

In this connection, the expression for the cross-talk attenuation of the capacitive-coupling currents takes the following form:

$$B_c = \ln\frac{I_1}{I_{2c}} = \ln\left|\frac{2}{(g + j\omega c)Z}\right|. \qquad (6\text{-}9)$$

The cross-talk attenuation of the inductive-coupling currents will be

$$B_m = \ln\frac{I_1}{I_{2m}} = \ln\left|\frac{2Z}{r + j\omega m}\right|. \qquad (6\text{-}10)$$

The expressions obtained establish the connection between the primary interaction parameters r, m, c, g, and the secondary parameter B, and permit calculation of the value of the cross-talk attenuation in short lengths of communications cables.

Let us consider the effects at the near and far ends of a cable circuit owing to the combined effect of the capacitive and inductive coupling currents.

In order to do this, we must first of all establish the law by which the capacitive and magnetic coupling coefficients may be added at the near and far ends of the cable.

As can be seen from Fig. 6-4, the noise current arising in the second circuit owing to capacitive coupling, $\dot{I}_{2c}$, splits parallel to the near and far ends of the cable circuit. The inductive coupling, acting as a transformer, gives rise to an inductive-noise current $\dot{I}_{2m}$, closed in series with the near and far ends of the second circuit.

As a result, two currents pass through the loads (equipment) at the near and far ends of the disturbed circuit; a capacitive-coupling current and an inductive-coupling current, where the sum of these currents acts at the near end, and their difference at the far end.

In receiving equipment of the second circuit, located at the same point as the transmitting equipment, the capacitive and inductive noise add ($\dot{I}_{20} = \dot{I}_{2c} + \dot{I}_{2m}$), while in receiving equipment located at the opposite end, the induced current is subtracted from the capacitive current ($\dot{I}_{21} = \dot{I}_{2c} - \dot{I}_{2m}$). This is confirmed experimentally by an investigation of the electromagnetic coupling coefficients carried out on shipping-length sections of cable.

Figure 6-5 shows the experimental set-up for studying electromagnetic coupling. The measurements of capacitive coupling are carried out with the ends of the line open-circuited; for the inductive-coupling measurements, they are short-circuited.

The law of addition of the capacitive and inductive interaction currents is found from the signs of the resultant coupling as measured at the near and far ends.

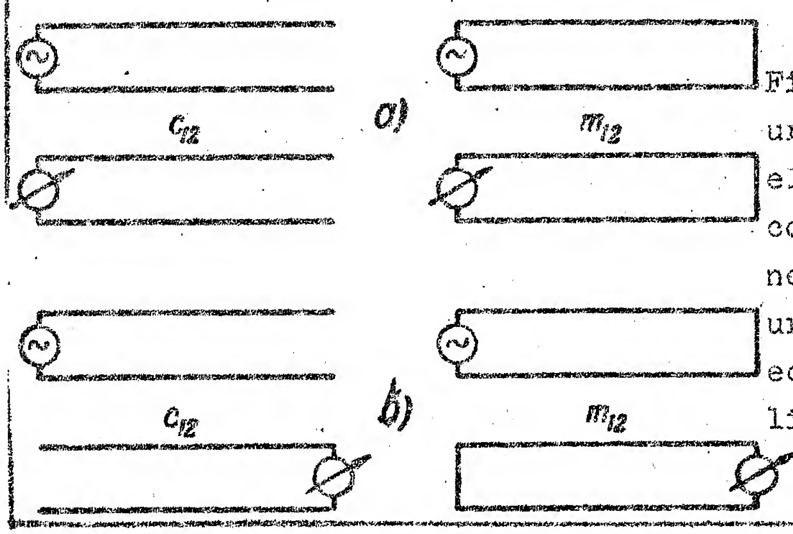


Fig. 6-5. a) Measurement of the electromagnetic coupling at the near end; b) Measurement of the electromagnetic coupling at far end.

The measurements established that the magnitude and sign of the capacitive coupling (C_{12}) were the same at the near and far ends. The inductive couplings (M_{12}) at the near and far ends differed in sign, but had the same absolute value.

The results of these measurements are related both to the reactive coupling components, c, m, and to g, r, the active components. Thus, at the near end, the inductive coupling is added to the capacitive ($C_{12} + M_{12}$), while at the far end, it is subtracted ($C_{12} - M_{12}$). At the near end, therefore, there acts the sum of the capacitive and inductive noise currents, while at the far end, their difference appears. Consequently, there will be less interference at the far end than at the near.

Using Formulas (6-7) and (6-8), we obtain the expressions for the resultant noise currents for the near end of the cable curcuit $\dot{I}_{20}$, and the far end, $\dot{I}_{21}$:

$$\dot{I}_{20} = \dot{I}_{2c} + \dot{I}_{2m} = \frac{1}{2}\left(g + j\omega c\right)\dot{I}_1 Z + \frac{1}{2}(r + j\omega m)\frac{\dot{I}_1}{Z} = \\ = \dot{I}_1\left[\frac{(g + j\omega c)Z^2 + (r + j\omega m)}{2Z}\right]. \qquad (6\text{-}11)$$

$$\dot{I}_{2l} = \dot{I}_{2c} - \dot{I}_{2m} = \frac{1}{2}(g + j\omega c)\dot{I}_1 Z - \frac{1}{2}(r + j\omega m)\frac{\dot{I}}{Z} = \\ = \dot{I}_1\left[\frac{(g + j\omega c)Z^2 - (r + j\omega m)}{2Z}\right]. \qquad (6\text{-}12)$$

The cross-talk attenuation corresponding to the resultant noise currents at the near end is

$$B_0 = \ln \frac{I_1}{I_{20}} = \ln \left| \frac{2Z}{(g + j\omega c) Z^2 + (r + j\omega m)} \right| \qquad (6\text{-}13)$$

and at the far end

$$B_l = \ln \frac{I_1}{I_{2l}} = \ln \left| \frac{2Z}{(g + j\omega c) Z^2 - (r + j\omega m)} \right|. \qquad (6\text{-}14)$$

After some manipulation, we obtain the formulas for the cross-talk attenuation in the following form:

$$B_0 = \ln \left| \frac{2}{\omega Z \left[\left(\frac{g}{\omega} + jc \right) + \frac{\left(\frac{r}{\omega} + jm \right)}{Z^2} \right]} \right|. \qquad (6\text{-}15)$$

$$B_l = \ln \left| \frac{2}{\omega Z \left[\left(\frac{g}{\omega} + jc \right) - \frac{\left(\frac{r}{\omega} + jm \right)}{Z^2} \right]} \right|. \qquad (6\text{-}16)$$

Or, letting K_0 and K_l stand for the expressions in brackets,

$$K_0 = \left| \left[\left(\frac{g}{\omega} + jc \right) + \frac{\left(\frac{r}{\omega} + jm \right)}{Z^2} \right] \right|$$

and

$$K_l = \left| \left[\left(\frac{g}{\omega} + jc \right) - \frac{\left(\frac{r}{\omega} + jm \right)}{Z^2} \right] \right|,$$

we obtain the final formula for the cross-talk attenuation

$$B_0 = \ln\left|\frac{2}{\omega Z K_0}\right|, \qquad (6\text{-}17)$$

$$B_l = \ln\left|\frac{2}{\omega Z K_l}\right|. \qquad (6\text{-}18)$$

The quantities K_0 and K_l are called the coefficients of electromagnetic coupling corresponding to the near and far ends of the cable circuit.

By using Formulas (6-17) and (6-18) it is possible to determine the resultants of the cross-talk attentuation for the capacitive and inductive noise currents in sections of cable on the order of shipping length. They indicate that the cross-talk attenuation is lower, the greater the frequency ω of the transmitted current, the wave impedance Z of the cable, and the coefficients K_0 and K_l of electromagnetic coupling.

6-3. PRIMARY INTERACTION PARAMETERS

Capacitive Coupling c -- is the result of asymmetric direct capacitances between the conductors of the disturbed and disturbing circuits.

Figure 6-6a shows the disturbing loop I (conductors 1-2) and the disturbed loop II (conductors 3-4). The direct capacitances between the conductors, c_{13}, c_{23}, c_{14}, c_{24}, form a bridge.

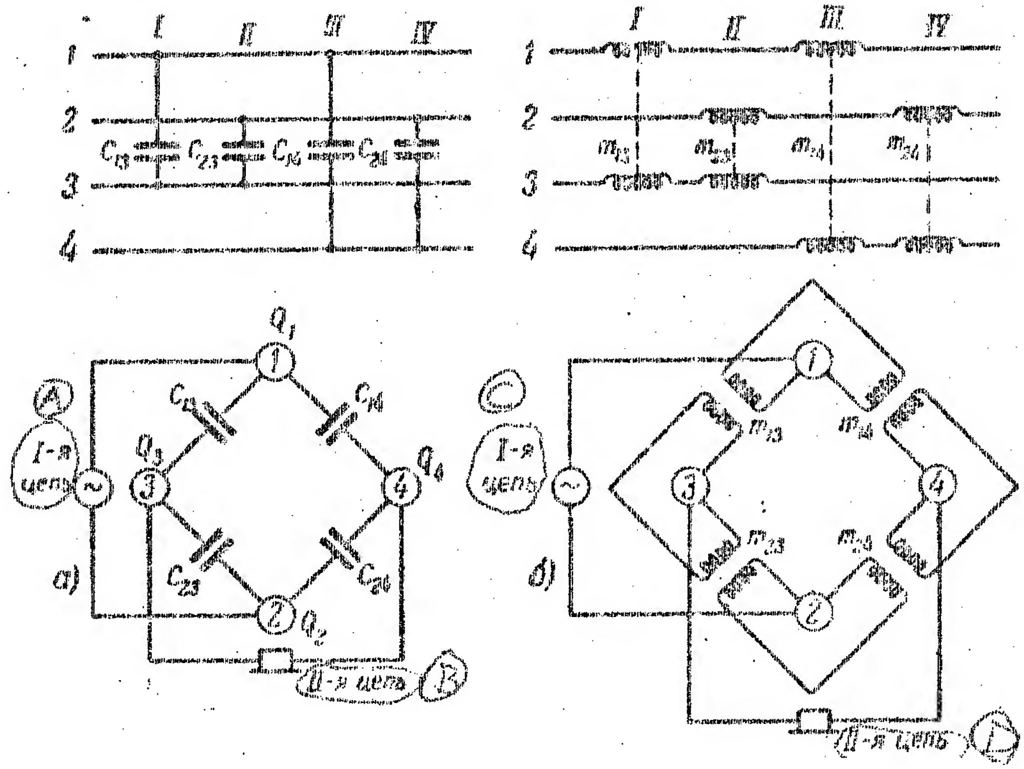


Fig. 6-6. Bridges formed by the direct capacitances and inductances of a quadded cable. A) First loop; B) second loop; C) first loop; D) second loop.

It is evident that if the bridge is symmetrical and balanced, there will be no transfer of energy (interference) between the first and second loops.

Let us explain the conditions for symmetry of the bridge.

On conductors 1-2 of loop I there are the electric-field charges Q_1 and Q_2, which induce the charges Q_3 and Q_4 on the conductors of loop II (3-4).

Each of the conductors 3 and 4 is affected by the difference of the charges Q_1 and Q_2 (which differ in sign):

$$Q_3 = Q_1 - Q_2 = \frac{U}{2} c_{13} - \frac{U}{2} c_{23},$$

$$Q_4 = Q_1 - Q_2 = \frac{U}{2} c_{14} - \frac{U}{2} c_{24}.$$

The charge difference acting on loop II equals:

$$Q_3 - Q_4 = \frac{U}{2}[(c_{13} - c_{23}) - (c_{14} - c_{24})] =$$

$$= \frac{U}{2}[(c_{13} + c_{24}) - (c_{14} + c_{23})].$$

It is evident that there will be no interference in loop II when the charge difference is zero:

$$Q_3 - Q_4 = 0,$$

and since $U/2 \neq 0$, this will be possible only where the direct capacitances are related by the equality:

$$(c_{13} + c_{24}) - (c_{14} + c_{23}) = 0,$$

which constitutes the condition for symmetry of the bridge.

Consequently, for the interaction between the loops to vanish, the sums of the opposing capacitances must be equal:

$$c_{13} + c_{24} = c_{14} + c_{23}.$$

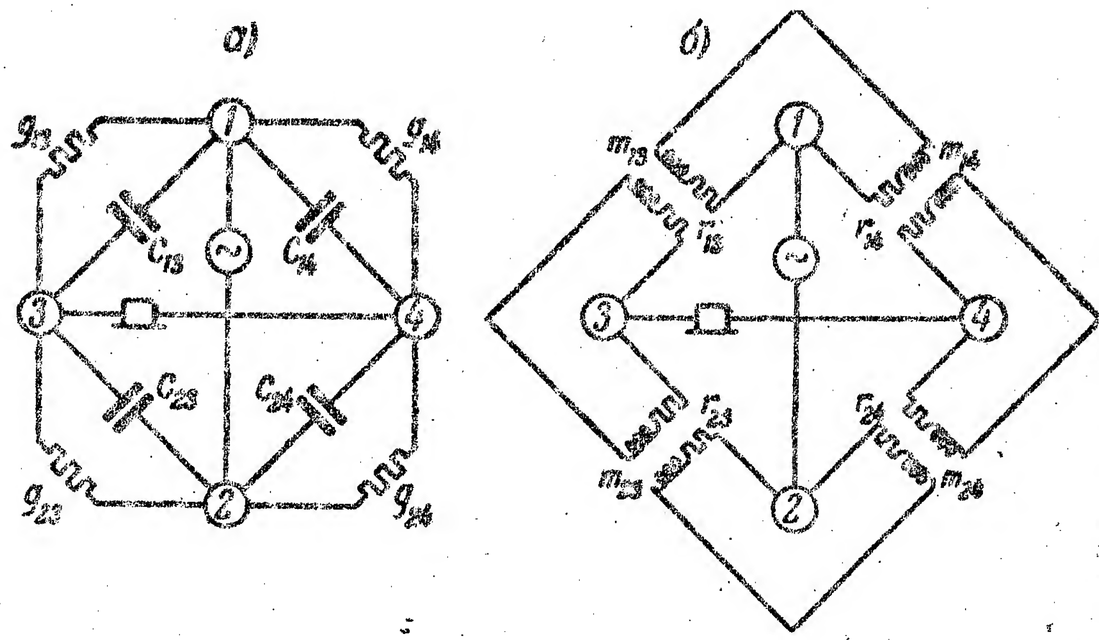


Fig. 6-7. Bridges formed by capacitive ($C_{12} = g + j\omega c$) and inductive coupling ($M_{12} = r + j\omega m$).

The capacitive asymmetry (unbalance) of the bridge which exists under actual conditions, causing mutual interference between communications circuits, is called the capacitive coupling c

$$c = (c_{13} + c_{24}) - (c_{14} + c_{23}).$$

Inductive coupling m -- can similarly be represented by a bridge formed by the direct inductances, which act as a transformer (Fig. 6-6b). Here we are concerned with magnetic fluxes rather than with electrical charges. The condition for symmetry of the bridge is the expression:

$$(m_{14} + m_{23}) - (m_{13} + m_{24}) = 0.$$

The inductive-coupling coefficient characterizes the tuning of the bridge, and accordingly the degree of energy transfer (interference) between loops I and II

$$m = (m_{14} + m_{23}) - (m_{13} + m_{34}).$$

The active component of the capacitive coupling or the dielectric coupling, g, is accounted for by the asymmetry of the energy loss in the dielectric. Here the arms of the bridge represent the equivalent energy losses in the dielectric surrounding the conductors of the cable, g_{13}, g_{23}, g_{24}, g_{14} (Fig. 6-7a).

When an alternating current flows through the cable, the dielectric introduces an energy loss proportional to

the conductance of the insulation, $G = \omega C \tan \delta$. If the electrical properties of the dielectric are not uniform, or the thickness of the conductor insulation varies, or the cable is deformed at various points, etc., then the direct losses in the dielectric g_{13}, g_{23}, g_{24}, and g_{14} will not be the same. This upsets the symmetry of the bridge formed by the dielectric couplings g, and sets up the conditions for mutual transfer of energy between the conductors.

The dielectric coupling is expressed by the equation:

$$g = (g_{13} + g_{24}) - (g_{14} - g_{23}).$$

The active component of the inductive coupling, or the so-called resistive galvanic coupling, r, is due to eddy currents.

As is known, when an alternating current is passed through a cable circuit, eddy currents are induced in adjacent conductors owing to the changing magnetic field; these currents cause an additional consumption of energy in the transmission circuit. Similar losses occur in the shield, lead sheath, and other metallic portions of the cable.

If the conductors of one circuit are not arranged symmetrically with respect to the conductors of another loop, or the metal sheaths of the cable, or if conductors

of differing diameters or electrical properties are used, the eddy-current losses will be asymmetric; this appears as a detuning of the resistive bridge r_{13}, r_{23}, r_{14}, and r_{24} (Fig. 6-7b), as illustrated in Fig. 6-8. Stronger eddy

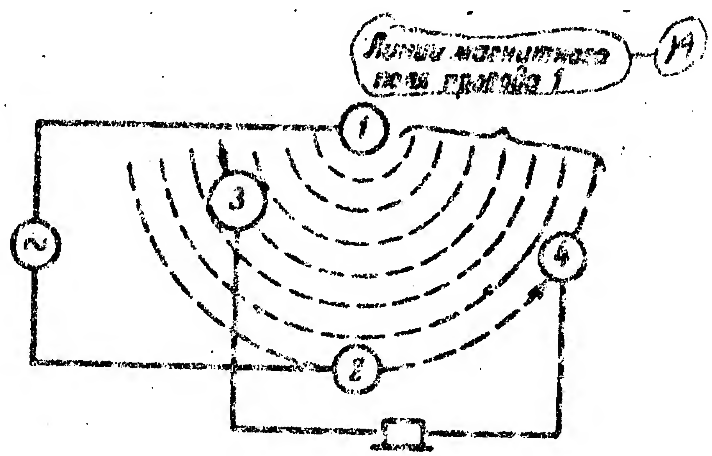


Fig. 6-8. Resistive coupling. 1-2) Distrubing circuit; 3-4) Disturbed circuit;
A) Magnetic lines of force due to conductor 1.

currents will be induced in the thick conductor 3, which lies near the disturbing loop 1-2, than in the thin conductor 4, located some distance from it. As a result, the resistive energy losses become asymmetric; this is characteristic of resistive coupling

$$r = (r_{14} + r_{23}) - (r_{13} + r_{24}).$$

The resistive coupling increases the more the conductors differ in resistance and in the losses to eddy currents in the adjacent circuit, shield, armoring, lead,

or other metal portions of the cable.

While asymmetric losses in the dielectric are responsible for dielectric coupling, asymmetric losses in metal cause resistive coupling.

It is evident that dielectric and resistive coupling occur only with changing magnetic fields. Inductive coupling also occurs only when an alternating current is passed through the line.

The magnitudes of the capacitive and inductive coupling are expressed in terms of the direct interaction parameters by the following equations:

$$C_{12} = g + j\omega c = [(g_{13} + j\omega c_{13}) + (g_{24} + j\omega c_{24})] - \\ - [(g_{14} + j\omega c_{14}) + (g_{23} + j\omega c_{23})], \quad (6\text{-}19)$$

$$M_{12} = r + j\omega m = [(r_{14} + j\omega m_{14}) + (r_{23} + j\omega m_{23})] - \\ - [(r_{13} + j\omega m_{13}) + (r_{24} + j\omega m_{24})]. \quad (6\text{-}20)$$

6-4. ELECTROMAGNETIC COUPLING COEFFICIENTS AND CROSS TALK ATTENUATION IN SHORT LENGTHS OF CABLE

The resultant coefficients of electromagnetic coupling at the near end, K_0, and the far end, K_1, may be represented as the sum of the electrical and magnetic couplings, K_c and K_m:

$$K_0 = K_c + K_m = \left(\frac{g}{\omega} + jc\right) + \frac{\left(\frac{r}{\omega} + jm\right)}{Z^2} = \frac{C_{12}}{\omega} + \frac{M_{12}}{\omega Z^2}. \quad (6\text{-}21)$$

$$K_l = K_c - K_m = \left(\frac{g}{\omega} + jc\right) - \frac{\left(\frac{r}{\omega} + jm\right)}{Z^2} = \frac{C_{12}}{\omega} - \frac{M_{12}}{\omega Z^2}. \quad (6\text{-}22)$$

Grouping the in-phase and out-of-phase components, we obtain:

$$K_0 = \left(\frac{g}{4\omega} + \frac{r}{\omega z^2}\right) + j\left(\frac{k}{4} + \frac{m}{z^2}\right).$$

$$K_l = \left(\frac{g}{4\omega} - \frac{r}{\omega z^2}\right) + j\left(\frac{k}{4} - \frac{m}{z^2}\right).$$

Bearing in mind that in cable measurements, the parameter k = 4c is used, rather than the capacitive coupling c, the electromagnetic coupling coefficient at the near and far ends can be expressed as:

$$K_0 = \left(\frac{g}{\omega} + \frac{r}{\omega z^2}\right) + j\left(c + \frac{m}{z^2}\right). \quad (6\text{-}23)$$

$$K_l = \left(\frac{g}{\omega} - \frac{r}{\omega z^2}\right) + j\left(c - \frac{m}{z^2}\right). \quad (6\text{-}24)$$

The absolute values of the quantities K_0 and K_l of expressions (6-21) and (6-22), occurring in the formulas used to determine the cross-talk attenuation B_0 and B_l, equal:

$$|K_0| = \sqrt{\left(\frac{g}{\omega} + \frac{r}{\omega z^2}\right)^2 + \left(c + \frac{m}{z^2}\right)^2}. \quad (6\text{-}25)$$

$$|K_l| = \sqrt{\left(\frac{g}{\omega} - \frac{r}{\omega z^2}\right)^2 + \left(c - \frac{m}{z^2}\right)^2}. \qquad (6\text{-}26)$$

When making calculations for an aerial line, the in-phase coupling components g/ω and $r/\omega z^2$ are neglected, giving:

$$|K_0| = c + \frac{m}{z^2}; \quad |K_l| = c - \frac{m}{z^2}.$$

In low-frequency cables, where the interaction is chiefly determined by the capacitive coupling, the coefficients $|K_0|$ and $|K_1|$ are equal:

$$|K_0| = |K_l| = c.$$

Here the cross-talk attenuation at the near and far ends of a shipping-length section of cable are also equal

$$B_0 = B_l = \ln\left|\frac{2}{\omega Z c}\right|. \qquad (6\text{-}27)$$

If k/4 is used instead of c, we obtain:

$$B_0 = B_l = \ln\left|\frac{8}{\omega Z k}\right|. \qquad (6\text{-}28)$$

The following units are used in measuring and calculating the electromagnetic-coupling coefficients:

$C_{12} = g + j\omega c$ [mhos]; $M_{12} = r + j\omega m$ [ohms];

$\frac{g}{\omega} + jc$ [farads]; $\frac{r}{\omega} + jm$ [henrys]; $\frac{\frac{r}{\omega} + jm}{z^2}$ [farads];

$K_0 = K_c + K_m$ [farads]; $K_1 = K_c - K_m$ [farads].

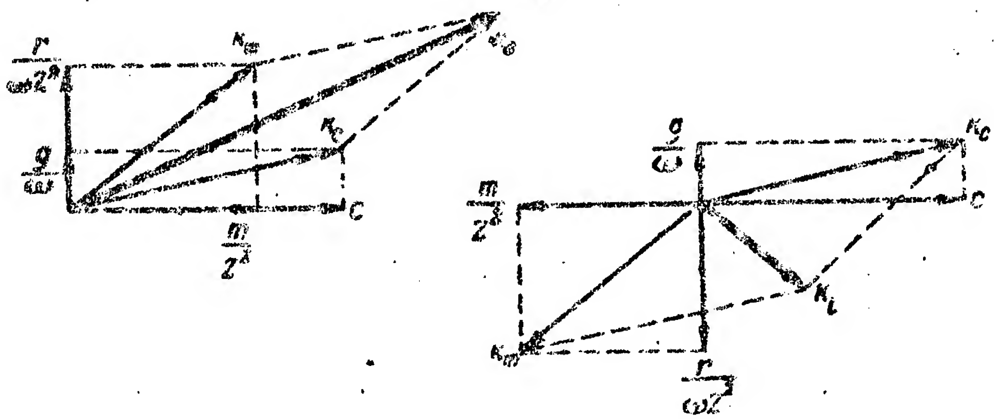


Fig. 6-9. Coupling vectors K_0 and K_1 in a quad (typical cases). a) at near end; b) at far end.

Normally, the values of K_0 and K_1 are on the order of 10^{-12} farad for a shipping length line cable.

Since in the general case coupling is expressed by complex vector quantities, K_1 and K_0 may differ in absolute value. If the signs of the capacitive and inductive components of coupling are the same, then $|K_0| > |K_1|$; where the signs differ, $|K_0| < |K_1|$.

Figure 6-9 shows the most typical cases of the coupling relationships inside a spiral quad.

Table 6-1 gives values of the K_0 and K_1 vectors inside quads of cables with styroflex-cord insulation for frequencies up to 60,000 cps.

Table 6-2 gives the mean values of a number of measured values of K_0 and K_1 inside and outside of quads of

type 32 X 2 paper-cord insulated cables.

Table 6-1. Coupling coefficients (in micromicrofarads) at the near and far ends of a cable (inside quads).

A) frequency, cps;   B) Specimen.

A) Частота, гц	B) I образец	
	K_0	K_l
1 000	0,7 ∠+100°	0,7 ∠+91°
10 000	1,1 ∠+113°	0,3 ∠+17°
20 000	1,38 ∠+102°	0,25 ∠−37°
60 000	1,7 ∠+94°	0,15 ∠−68°
A) Частота, гц	B) II образец	
	K_0	K_l
1 000	0,25 ∠−63°	0,2 ∠−79°
10 000	0,53 ∠−90°	0,16 ∠+14°
20 000	0,35 ∠−92°	0,09 ∠0°
60 000	0,42 ∠−104°	0,06 ∠0°
A) Частота, гц	B) III образец	
	K_0	K_l
1 000	2,2 ∠+95°	2,8 ∠+96°
10 000	4,5 ∠+108°	1,7 ∠+42°
20 000	5,4 ∠+102°	1,0 ∠+18°
60 000	6,8 ∠+106°	1,1 ∠0°

From the data given, it follows that for intersections within the quad, the capacitive and inductive

coupling are the same in the vast majority of cases, since the coefficient of electromagnetic coupling at the far end, $|K_1|$ is much less than the coefficient of electromagnetic coupling at the near end, $|K_0|$.

Between the quads, the capacitive and inductive coupling agree or disagree in sign, but here $K_0 > K_1$.

Table 6-2. Coefficients K_0 and K_1 for a 32 X 2 cable. A) f, cps; B) inside quads; C) between quads

(A) f, кгц	(B) Внутри четверок		(C) Между четверками	
	$\|K_0\|$	$\|K_1\|$	$\|K_0\|$	$\|K_1\|$
5	11,0	1,5	0,5	0,8
10	12,6	0,7	0,2	1,0
20	13,5	0,75	0,1	1,1
30	13,9	1,0	0,2	1,1
40	14,3	1,4	0,4	1,0
50	15,4	1,8	0,7	0,9
60	17,7	2,1	1,0	0,9
70	20,1	3,1	1,5	0,89
80	22,5	4,6	2,0	0,7

The coupling coefficients inside the quads are greater in absolute value than those between the quads, and consequently, the most dangerous interaction occurs inside the quads. This is confirmed by the results of measurements of the cross-talk attenuation in shipping-length type 32 X 2 paper-cord insulated cable (Table 6-3) and type

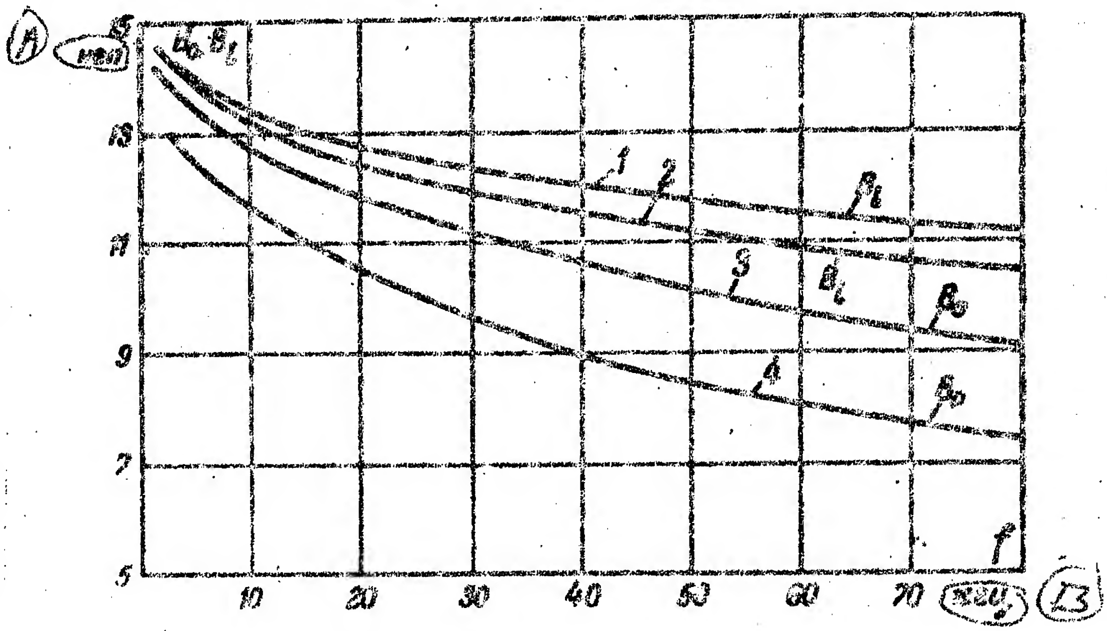


Fig. 6-10. Cross-talk attenuation B_0 and B_1 in a shipping length cable between and inside the quads (typical case). 1) Between the quads; 2) inside the quads; 3) between the quads; 4) inside the quads; A) nepers; B) kc.

4 X 4 styroflex-cord insulated cable (Fig. 6-10) .

It follows from Table 6-3 and Figure 6-10 that: 1. The cross-talk attenuation between quads is 1 to 4 nepers greater than the cross-talk attenuation inside the quads. Therefore transmissions through differing quads will be propagated under more favorable conditions than transmissions passing through loops within one quad. 2. The cross-talk attenuation B_1 at the far end is noticeably greater than the cross-talk attenuation B_0 at the near end; this effect is especially pronounced within the quads. 3. As the frequency increases, the cross-talk attenuation decreases considerably. Using formulas (6-17) and (6-18), the following relationships can be derived between the cross-talk attenuation and the coupling coefficient at the near and far ends of the cable:

$$B_l - B_0 = \ln\left|\frac{2}{\omega Z K_l}\right| - \ln\left|\frac{2}{\omega Z K_0}\right| = \ln\left|\frac{K_0}{K_l}\right|, \quad (6\text{-}29)$$

which agrees well with the experimental data. A coupling ratio between quads of $\left| K_0 / K_1 \right| = 2$ to 4 corresponds to B_1 exceeding B_0 by 0.5 to 1 neper.

Table 6-3. Results of measurements of cross-talk attenuation in shipping-length 32 X 2 cables.
A) frequency, kc: B) inside quads;
C) between quads

Частота, кгц (A)	(B) Внутри четверок							
	B_0	B_l	B_0	B_l	B_0	B_l	B_0	B_l
0,8	11,4	15	11,1	13,3	11,5	15	13,0	15
5,0	9,5	12,2	9,2	11,3	9,3	14,7	11,0	13,3
10,0	8,8	12,2	8,4	11	9,4	13	10,3	12,4
13,5	8,3	12,5	8,1	10,7	9,1	12,9	9,8	12,3
18	8,0	12,8	7,9	10,1	8,8	12,3	9,4	12,4
25	7,9	12,2	7,5	9,8	8,4	12,7	9,0	12,3
30	7,7	11,9	7,3	10	8,3	12,3	8,6	11,2
40	7,6	11,5	7,2	10	8,1	12,1	8,2	11,3
50	7,2	11,1	7,0	9,3	7,9	12,0	7,8	11,3
60	7,0	10,7	6,9	9,2	7,8	11,3	7,5	11
80	6,8	9,9	6,7	9,0	7,6	10,5	7,0	10,8

Частота, кгц (A)	(C) Между четверками							
	B_0	B_l	B_0	B_l	B_0	B_l	B_0	B_l
0,8	14,1	15	15	15	15	15	15	15
5,0	12,7	13,6	13,3	14,8	15	14,8	13,2	13,3
10,0	12,0	12,9	12,5	13,1	14,1	13,8	12,6	12,7
13,5	11,9	12,7	12,4	13,1	13,8	13,7	12,3	12,4
18	11,4	12,5	12,2	12,5	13,7	13,5	12,1	12,2
25	11,2	12,3	12,0	12,1	13,5	13,2	11,9	12,0
30	11,1	12,2	11,9	11,8	13,2	12,8	11,7	11,8
40	11,0	11,8	11,4	11,4	12,9	12,7	11,6	11,7
50	10,6	10,9	11,0	11,2	12,4	12,6	11,4	11,5
60	10,5	11,3	10,6	11,0	12,2	12,3	11,4	11,3
80	10,2	10,9	10,1	10,7	11,8	11,9	11,0	11,1

Table 6-4. Results of measurements of capacitive coupling in 32 X 2 cable, in micromicrofarads per standard (shipping) length. A) Frequency, kc; B) Inside first quad; C) Inside second quad; D) Inside third quad; E) Inside fourth quad; F) Between quads.

Часто-та, кгц	Внутри 1-й четверки k_1		Внутри 2-й четверки k_1		Внутри 3-й четверки k_1	
	k	$\frac{g}{\omega}$	k	$\frac{g}{\omega}$	k	$\frac{g}{\omega}$
0,8	+47,1	−0	—	—	+25,0	0
5,0	+49,0	−1,0	+69,0	−0,5	+25,0	0
10,0	+44,0	−1,4	+69,5	−1,0	+25,5	−0,25
13,5	+49,5	−2,0	+4[illegible],0	−2,0	+26,0	−1,5
18	+51,0	−2,0	+71,0	−2,25	+25,5	−1,0
25	+53,0	−2,5	+72,0	−2,75	+26,0	−1,5
30	+50,0	−2,5	−2,5	−2,5	+26,5	−1,0
40	+62,5	−3,0	+76,0	−3,0	+27,0	−2,0
50	+74,0	−2,0	+80,0	−2,5	+28,0	−2,0
60	+90,0	+1,0	+86,0	−2,0	+30,0	−2,0
80	120	+3,0	+111,0	+9,0	+37,0	+2,0

Внутри 4-й четверки k_1		Между четверками k_9		Между четверками k_{10}	
k	$\frac{g}{\omega}$	k	$\frac{g}{\omega}$	k	$\frac{g}{\omega}$
−8	0	−1,0	0	−2,0	0
−8	+0,2	−1,0	+1	−2,0	−0
−7,5	+0,8	−1	+0,25	−2,0	0
−7,1	+0,4	−1,25	+0,25	−2,0	+0,25
−6,1	+0,6	−1,25	+0,25	−2,0	+0,5
−3,4	+0,8	−1,0	+0,25	−2,0	+0,5
−1,0	+2,0	−1,0	+0,25	−2,0	+0,5
+5,6	+4,3	−0,5	+1,0	−1,5	+1,0
+16	+4,0	+0,5	+1,0	−1,0	+1,0
+30,0	+13	+1,5	+1,0	0	+2,0
+107,0	+3,5	+4,5	+2,0	+2,0	+2,5

Table 6-5. Results of measurement of magnetic couplings in 32 X 2 cable (in millimicrohenrys per standard (shipping) length). A) Frequency, kc; B) Inside first quad; C) Inside second quad; D) Inside third quad; E) Inside fourth quad; F) Between quads.

A Часто-та, кгц	B Внутри 1-й четверки m_1: m	B Внутри 1-й четверки m_1: $\frac{r}{\omega}$	C Внутри 2-й четверки m_1: m	C Внутри 2-й четверки m_1: $\frac{r}{\omega}$	D Внутри 3-й четверки m_1: m	D Внутри 3-й четверки m_1: $\frac{r}{\omega}$
0,8	+475	+30	+560	+30	+265	+15
5,0	+475	+15	+555	+15	+265	+15
10,0	+480	+20	+560	+15	+265	+20
13,5	+495	+20	+575	+10	+265	+20
18	+500	+20	+570	+10	+260	+20
95	+530	+20	+580	+5	+260	+20
30	+560	+30	+600	+2	+235	+20
40	+630	+45	+620	—2	+240	+20
50	+750	+50	+675	—5	+300	+20
60	+940	+11,0	+775	0	+330	+20
80	1 200	—	1 200	—	+490	+35

A Часто-та, кгц	E Внутри 4-й четверки m_1: m	E Внутри 4-й четверки m_1: $\frac{r}{\omega}$	F Между четверками m_9: m	F Между четверками m_9: $\frac{r}{\omega}$	F Между четверками m_{10}: m	F Между четверками m_{10}: $\frac{r}{\omega}$
0,8	+95	0	—40	0	—15	—5
5,0	—95	0	—30	—10	—10	0
10,0	—95	+10	—30	0	—10	—10
13,5	—80	+15	—30	0	—10	—15
18	—70	+20	—30	0	0	—15
95	—55	+20	—35	0	+5	—10
30	—40	+30	—30	—2	+10	—10
40	+20	+60	—30	—5	+20	—10
50	+90	+75	—30	—5	+25	0
60	+210	+160	—30	0	+30	+5
80	+800	+550	—15	+10	+40	+30

Table 6-6. Percentage relationship of intraquad couplings. A) Frequency, kc; B) Inside first quad; C) Inside second quad; D) Inside third quad.

A Частота, кгц	B Внутри 1-й четверки				C Внутри 2-й четверки			
	$\frac{g}{\omega}$	c	$\frac{r}{\omega Z^2}$	$\frac{m}{Z^2}$	$\frac{g}{\omega}$	c	$\frac{r}{\omega Z^2}$	$\frac{m}{Z^2}$
0,8	0	85	1,4	13,6	0	84,2	0	15,8
5,0	0	53,2	2,6	44,2	2,2	89,5	0	8,3
13,5	1,2	46,6	2,7	48,5	6,7	42,4	1,9	49,0
30	1,76	45	3,74	49,5	18,7	10,0	30,4	40,9
50	3,17	42,3	3,43	51,1	10,3	41,4	21,9	26,4
60	2,64	42	2,66	52,7	13,8	37,5	18,9	29,3
80	2,02	36,6	5,83	55,5	2,5	41,1	22,7	33,7

D Внутри 3-й четверки			
$\frac{g}{\omega}$	c	$\frac{r}{\omega Z^2}$	$\frac{m}{Z^2}$
0	85,5	1,05	13,45
1,07	55,5	1,33	42,1
1,42	48	1,58	49
1,62	45,5	2,68	50,2
1,2	44,4	3,4	51
0,3	48,2	5,0	51,5
—	—	—	—

6-5. THE FREQUENCY DEPENDENCE OF THE COEFFICIENTS OF ELECTROMAGNETIC COUPLING

The components of the electromagnetic coupling were measured in the 60 - 80 kc band for cables with paper-cord (type 32 X 2) and styroflex-cord (type 4 X 4) insulation; then, using expressions (6-21) and (6-22), the values of K_0 and K_1 were calculated.

Tables 6-4 and 6-5 give the results of measurements of the capacitive and inductive coupling inside and between the quads of 32 X 2 paper-cord insulated cable.

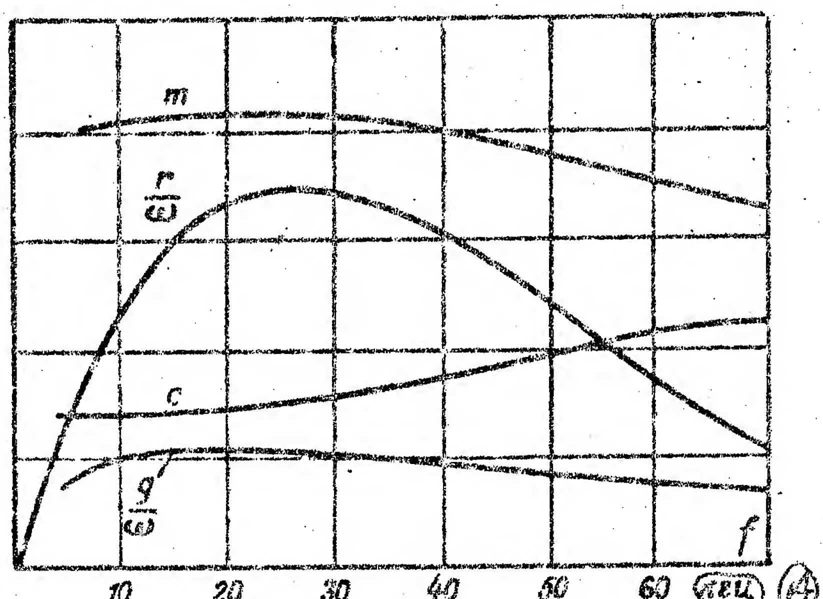


Fig. 6-11. Nature of the frequency variation of the coupling coefficients. A) kc.

The nature of the dependence of the resistive, in-

ductive, capacitive, and dielectric couplings on frequency is shown in Fig. 6-11, for frequencies up to 70,000 cps.

The relative importance of the various components of coupling in creating interference between circuits is illustrated by their percentage relationship, given for 32 X 2 cable in Tables 6-6 and 6-7.

Table 6-7. Percentage relationship of interquad couplings. A) Frequency, kc

Частота, кгц (A)	g/ω	c	$r/\omega Z^2$	m/Z^2	g/ω	c	$r/\omega Z^2$	m/Z^2
0,8	0	80	7,5	12,5	0	61,5	0	38,5
5,0	0	39	30,5	30,5	19,5	19,5	15,3	45,7
13,5	3,5	36,4	13,7	46,4	4	28	0	68
30	9,55	38,2	10,45	41,8	5,35	21	4,52	69
50	16,75	16,75	18,9	47,5	18,3	9,3	10,4	62
60	30,4	0	8,6	61	14,6	26	0	59,4
80	21,5	17,9	19,6	41	21,5	51,7	8,3	18,5

Figure 12 shows the nature of the frequency variation in the percentage relationship of the couplings in a type 4 X 4 cable with styroflex-cord insulation. The frequency dependence of the ratio $\left| K_c / K_m \right|$ for a cable with styroflex insulation is given in Fig. 6-13.

The typical frequency variation of $|K_0|$ and $|K_1|$ is shown in Fig. 6-14.

Analyzing the data which has been presented, we see that:

1. At voice frequencies, the capacitive coupling is six to twelve times the inductive coupling. As the fre-

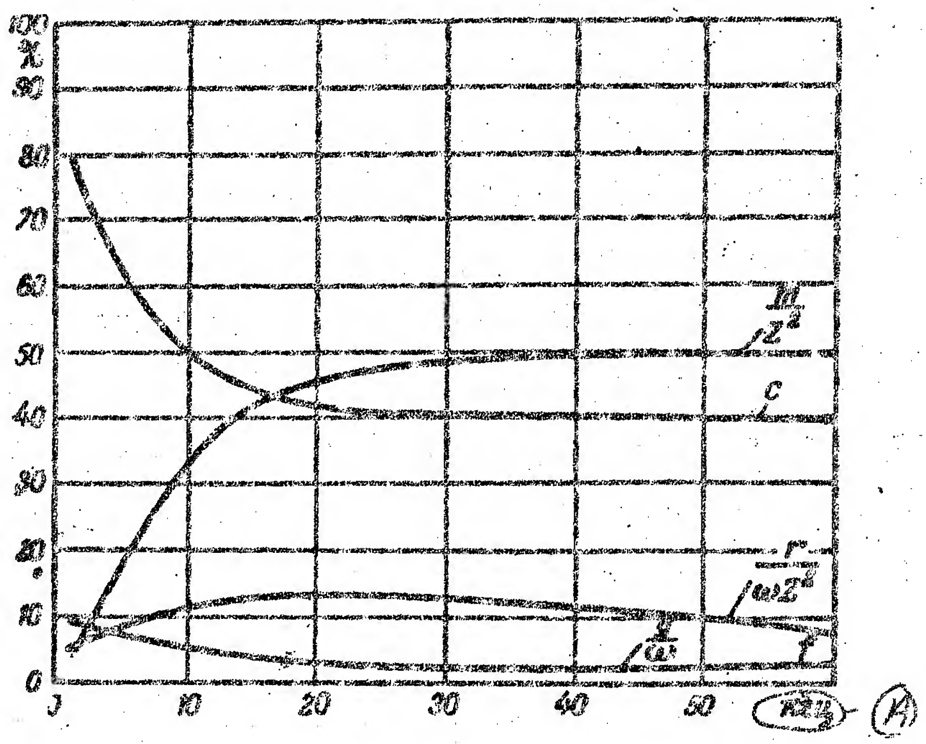


Fig. 12. The percentage relationship of intre-quad coupling in the cable (typical case). A) kc.

quency increases, however, the relationship changes, and by the time f reaches 10,000 cps, the inductive coupling is numerically equal to the capacitive coupling, and then somewhat exceeds it (in several quads, the ratio $|K_c / K_m|$ reached 0.7 to 0.9 at frequencies of 60 to 80 kc). On the average, from 10 kc on K_c / K_m equalled approximately one.

2. The in-phase coupling components, especially the resistive component, play a more important rôle as the frequency increases (with DC, they are zero).

Thus, although in the voice-frequency range r/ω amounts to only 4 to 8% of the magnitude of the inductive couplings, at higher frequencies this percentage rises to 20%.

The in-phase component of the capacitive coupling (the dielectric coupling g/ω) is relatively unimportant, especially in cables with styroflex insulation, where the dielectric losses are insignificant, owing to the small value of tan δ.

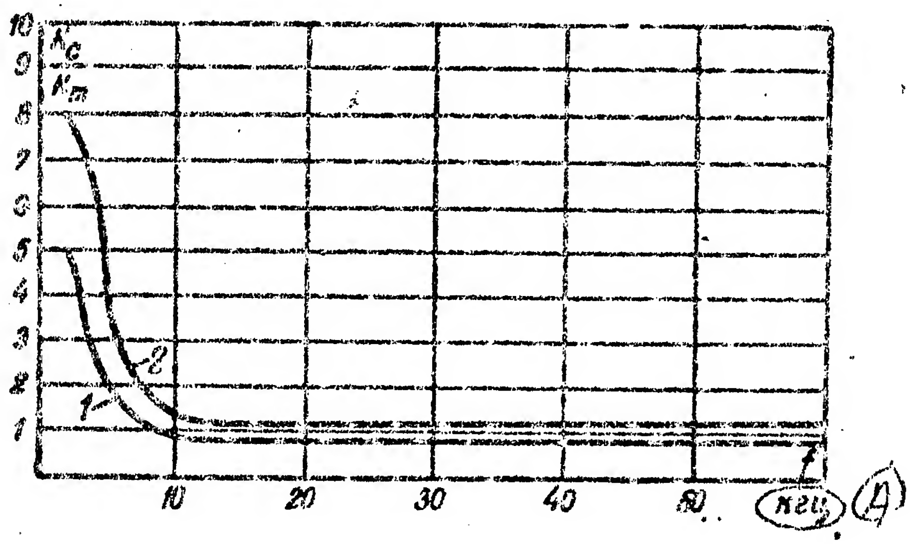


Fig. 6-13. The dependence of the relationship of the capacitive and inductive couplings on frequency (typical case). 1) intraquad; 2) interquad; A) kc.

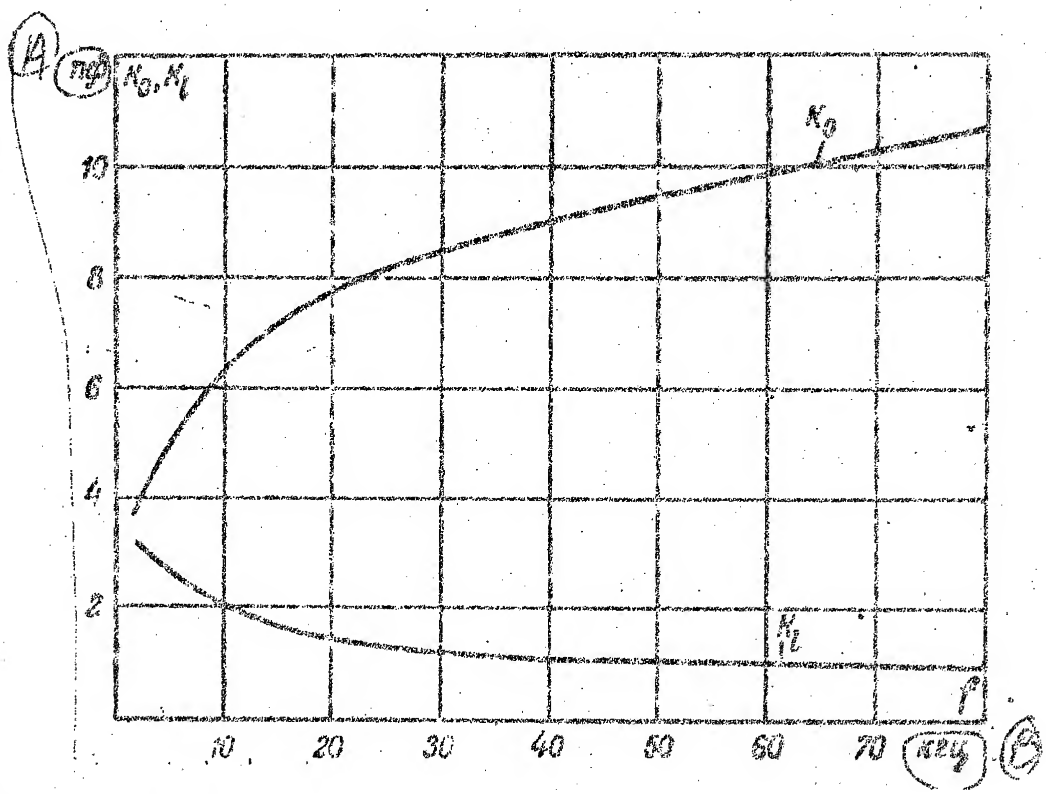


Fig. 6-14. Frequency dependence of intraquad electromagnetic coupling coefficients (typical case). A) micromicrofarads; B) kc.

The mean value of the ratio of the in-phase and out-of-phase coupling components, measured for several specimens, equals:

$$\frac{g}{\omega} : c = \quad 0.1 \text{ to } 0.2,$$

$$\frac{r}{\omega} : m = 0.2 \text{ to } 0.4.$$

3. Frequency dependence of the coupling components:

a) As the frequency increases, the capacitive coupling becomes somewhat greater; this is especially true of cables with paper-cord insulation. This change is evidently connected with the variation of the dielectric constant with frequency.

b) At frequencies of f = 10 to 20 kc, the inductive coupling begins to decrease uniformly; this is explained by a proximity effect, both for the circuits measured, and for adjacent quads and other metal portions of the cable (the lead sheath, etc.).

c) The resistive coupling r/ω rises from zero (for DC), reaching a maximum at frequencies of 15 to 30 kc, and then falls.

Physically, this is explained by the fact that for very large eddy-current losses asymmetry (i.e., resistive coupling) is unimportant.

d) Dielectric coupling, g/ω, also comes into play only for AC, and reaches a maximum at a particular frequency. In styroflex-insulated cables, this maximum falls in the voice-frequency band, while in paper-cord--insulated cables, it occurs at 30 to 40 kc.

All the components of coupling must be considered in order to meet the noise rejection standards for long-

distanse high-frequency multiplexed cables.

Where cables are to be used only in the low-frequency band (up to 3,000 cps), only the capacitive coupling coefficients need be known (for DC, m, r, and g have no effect).

The frequency dependence of the near-end $|K_0|$ and far end $|K_1|$ electromagnetic coupling coefficients (Fig. 6-14) is determined by the fact that at low frequencies, where interaction results from capacitive asymmetry alone, K_0 and K_1 are nearly equal. As the frequency increases, the importance of the inductive couplings rises, and the in-phase components r and g appear. In view of the fact that at the near end, the capacitive coupling K_c and the inductive coupling K_m add, the coefficient K_0 must increase. At the far end, the inductive coupling subtracts from the capacitive, so that the couplings compensate each other, and the coefficient K_1 decreases; thus in the high frequency band K_0 has an absolute value 5 to 20 times that of K_1. This is typical of the interaction of the conductors of a quad.

6-6. ELECTROMAGNETIC COUPLING BETWEEN ANY CIRCUITS IN A CABLE

Above, we have considered the interaction of two circuits lying within one spiral quad of a cable.

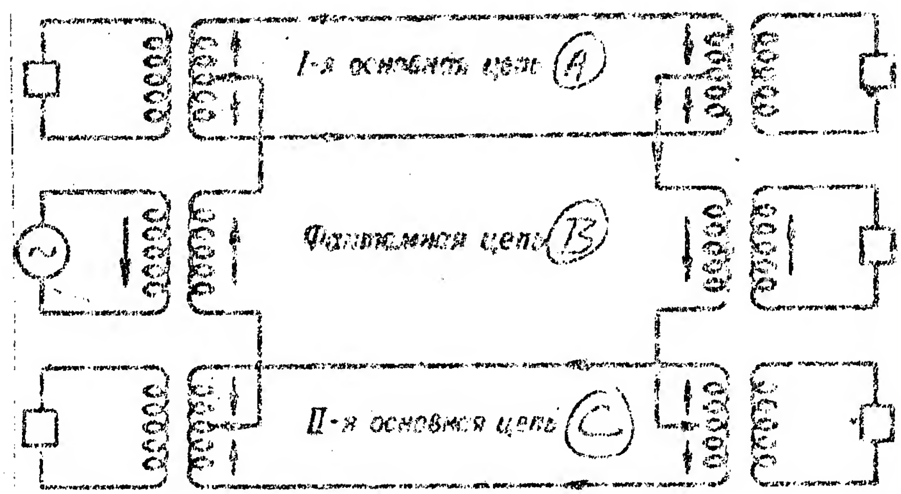


Fig. 6-15. Phantom circuit: A) First physical circuit; B) phantom circuit; C) second physical circuit.

It is quite natural that when energy is transmitted over a multiconductor cable, loops lying in different quads will interact.

The phantom circuit (Fig. 6-15), used in low-frequency cables, permits a single quad to be used for not two, but three communications circuits; this leads to interaction between the physical and phantom circuits. Owing to the direct capacitances of the current-bearing conductors and the lead sheath, there is an interaction through the ground.

Each form of interaction between loops has been given an appropriate index.

Table 6-8 gives the generally accepted designations

for the coupling coefficients of different circuits. In Fig. 6-16, the loops in two quads are shown.

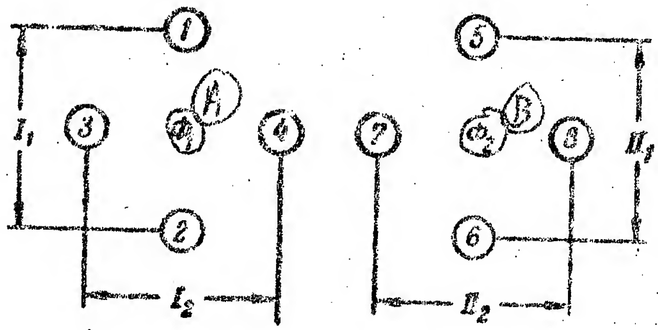


Fig. 6-16. Designating the coupling coefficients between quads. A) Ph_1; B) Ph_2.

Table 6-8. Designations of coupling coefficients between different circuits.

COEFFICIENT	CIRCUITS INVOLVED	DESIGNATION
k_1	I physical/II physical	I/II
k_2	I physical/phantom	I/Ph
k_3	II physical/phantom	II/Ph
e_1	I physical/ground	I/Gr
e_2	II physical/ground	II/Gr
e_3	Phantom/ground	Ph/Gr
k_4	Phantom/phantom	Ph_1/Ph_2
k_5	I physical-I quad/phantom-II quad	I_1/Ph_2
k_6	II physical-I quad/phantom-II quad	II_1/Ph_2
k_7	Phantom-quad I/I physical-quad II	Ph_1/I_2
k_8	Phantom-I quad/ II physical-II quad	Ph_1/II_2
k_9	I physical-I quad/II physical-II quad	I_1/II_2
k_{10}	I physical-I quad/I physical-II quad	I_1/I_2
k_{11}	II physical-I quad/I physical-II quad	II_1/I_2

Table 6-8 (continued).

COEFFICIENT	CIRCUITS INVOLVED	DESIGNATION
k_{12}	II physical-I quad/II physical-II quad	II_1/II_2

It is clear from Table 6-8 that k and e with indices 1 to 3 determine the interactions within a single quad. The coefficients with the indices 4 to 12 characterize interaction between individual loops of two different quads. To distinguish it from k, which is called the capacitive coupling coefficient, e is called the capacitive unbalance coefficient.

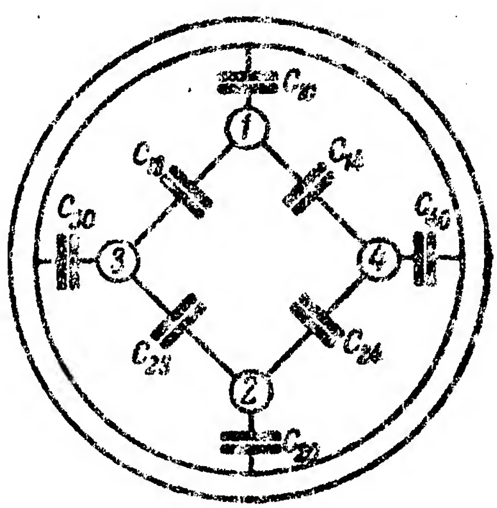


Fig. 6-17. Direct capacitances of a quad.

Table 6-9 gives all the values of all the capacitive coupling and asymmetry coefficients that involve the direct capacitances (the indices denote the number of the corresponding conductor).

Figure 6-17 gives a diagram of the direct capacitances within a quad; Fig. 6-18, the same between quads.

The coefficient k is related mathematically to the previously derived values for the capacitive coupling c by the following relationships

$$c = k_4;\quad c = \frac{k_1}{4};\quad c = \frac{k_{9-12}}{4};\quad c = \frac{k_{2-3}}{2};\quad c = \frac{k_{5-8}}{2}.$$

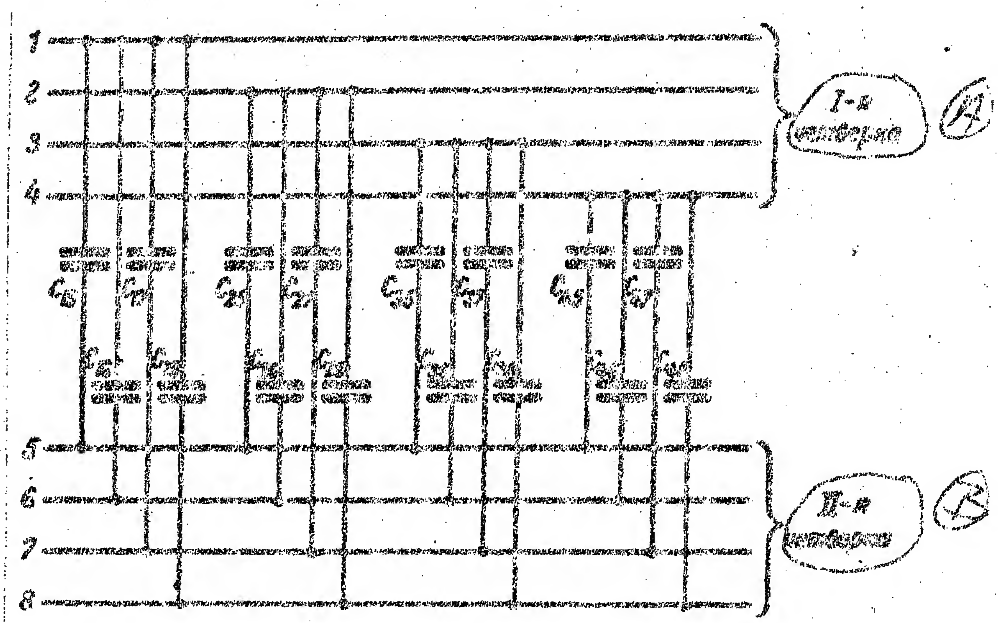


Fig. 6-18. Direct capacitances between two quads. A) First quad; B) second quad.

As was shown above, it is only necessary to compute

the capacitive coupling for a low-frequency cable.

Table 6-9. Values of the capacitive coupling and asymmetry coefficients in cables. A) Values of the coefficients.

A) Значения коэффициентов

$$k_1 = (c_{13} + c_{24}) - (c_{14} + c_{23})$$
$$k_2 = (c_{13} + c_{14}) - (c_{23} + c_{24})$$
$$k_3 = (c_{13} + c_{23}) - (c_{14} + c_{24})$$
$$e_1 = (c_{10} - c_{20})$$
$$e_2 = (c_{30} - c_{40})$$
$$e_3 = (c_{10} + c_{20}) - (c_{30} + c_{40})$$
$$k_4 = (c_{15} + c_{16} + c_{25} + c_{26} + c_{37} + c_{38} + c_{47} + c_{48}) - (c_{17} + c_{18} + c_{27} + c_{28} + c_{35} + c_{36} + c_{45} + c_{46})$$
$$k_5 = (c_{15} + c_{16} + c_{27} + c_{28}) - (c_{17} + c_{18} + c_{25} + c_{26})$$
$$k_6 = (c_{35} + c_{36} + c_{47} + c_{48}) - (c_{37} + c_{38} + c_{45} + c_{46})$$
$$k_7 = (c_{15} + c_{25} + c_{36} + c_{46}) - (c_{16} + c_{26} + c_{35} + c_{45})$$
$$k_8 = (c_{17} + c_{27} + c_{38} + c_{48}) - (c_{18} + c_{28} + c_{37} + c_{47})$$
$$k_9 = (c_{15} + c_{26}) - (c_{16} + c_{25})$$
$$k_{10} = (c_{17} + c_{28}) - (c_{18} + c_{27})$$
$$k_{11} = (c_{35} + c_{46}) - (c_{36} + c_{45})$$
$$k_{12} = (c_{37} + c_{48}) - (c_{38} + c_{47})$$

In high-frequency cables, the interaction is determined by the capacitive and inductive couplings acting together. In this case, therefore, it is necessary to use the quantities k and m. In the designations for the inductive coupling, analogous indices are used; since high-frequency loops do not use phantom circuits, in calculations for HF cables, only k_1 and k_9 through k_{12} are used in addition to m_1 and m_9 through m_{12}.

6-7 CROSS-TALK ATTENUATION IN LONG CABLE LINKS

The concepts and calculational formulas presented above are relevant to short sections of a cable circuit

on the order of shipping length.

For practical purposes, it is important to establish the laws for the interaction and the magnitudes of the cross-talk attenuations in long lines.

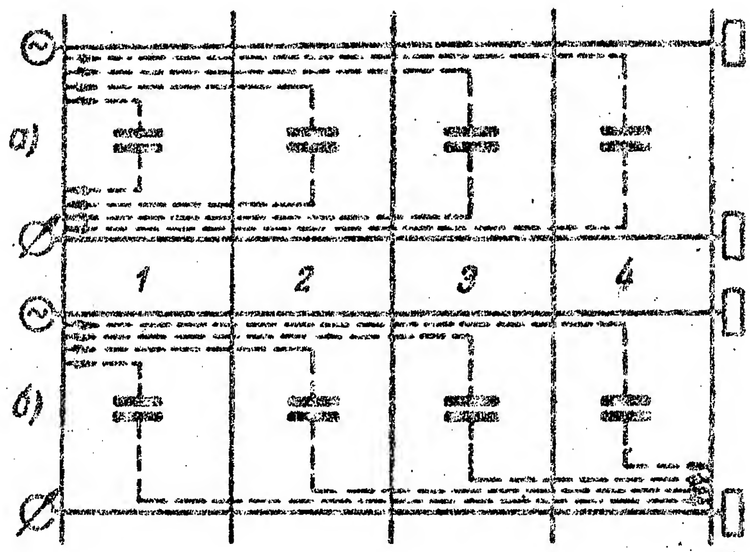


Fig. 6-19. Interaction of cable circuits several shipping-length sections long. a) Near-end effect; b) far-end effect.

At the present time, the interaction effect in a long cable line is assumed to equal the geometric sum of the interactions of its individual sections. The use of the geometric, rather than the arithmetic, sum explains why for cable lines, in contrast to aerial lines, the phases of the interaction currents leaving the various sections of the cables are not known.

Thus, adding the transferred currents of the individual cable sections geometrically, we obtain:

$$I_{tot} = \sqrt{I_1^2 + I_2^2 + \cdots + I_n^2} \,.$$

The difference in the addition of the interaction currents at the near and far ends is illustrated in Fig. 6.19.

A consideration of the interaction at the far end shows that the current elements flowing from each shipping-length section of the cable traverse practically the same path and are identically attenuated. For interaction at the near end, the currents leaving the various cable sections are not the same, since the further from the input the cable section is located, the greater the attenuation of the current leaving each section.

The resultant interference-current element I^k (due to capacitive and inductive coupling), flowing from the k-th section of the line to the input of the circuit, causing an interaction, may be represented at the end of the line as:

$$I_{20}^k = I_1 e^{-2\beta s(n-1)} e^{-B_0},$$

where I_1 is the current at the input of the disturbing loop;

$2\beta s(n-1)$ is the attenuation of the first and second loops themselves over the distance from the input to the k-th section of the line;

B_0 is the cross-talk attenuation between the first and second circuits for a standard (shipping) length of cable;

β is the linear attenuation, nepers/km;

s is the shipping length of the cable, km;

n is the total number of sections.

Using the laws for the geometric addition of elements of current, we obtain the total interference current at the near end of the second circuit

$$I_{20} = \sqrt{\sum_{n=1}^{n} (I_{20}^{k})^2} = \sqrt{\sum_{n=1}^{n} (I_1 e^{-2\beta s(n-1)} e^{-B_0})^2} =$$

$$= I_1 e^{-B_0} \sqrt{\sum_{n=1}^{n} e^{-4\beta s(n-1)}}.$$

Hence the cross-talk attenuation for a long cable at the near end is

$$B_{0n} = \ln \frac{I_1}{I_{20}} = \ln \frac{I_1}{I_1 e^{-B_0} \sqrt{\sum_{n=1}^{n} e^{-4\beta s(n-1)}}} =$$

$$= B_0 - \ln \sqrt{\sum_{n=1}^{n} e^{-4\beta s(n-1)}}. \qquad (6\text{-}30)$$

On the basis of the law of geometrical progressions it is possible to represent the second term of (6-30) as:

$$\sqrt{\Sigma e^{-4\beta s(n-1)}} = \sqrt{\frac{1-e^{-4\beta ns}}{1-e^{-4\beta s}}} = \sqrt{\frac{\operatorname{sh} 2\beta ns}{\operatorname{sh} 2\beta s}} \sqrt{\frac{e^{2\beta s}}{e^{2\beta ns}}}$$

then, disregarding the quantity βs as negligible in comparison with βns, we obtain:

$$B_{0n} = B_0 - \ln \sqrt{\frac{\operatorname{sh} 2\beta ns}{\operatorname{sh} 2\beta s}} + \beta ns, \tag{6-31}$$

where βns is the attenuation of the entire cable line itself.

It is not hard to show that for a short line, where 2βns < 0.2

$$B_{0n} = B_0 - \ln \sqrt{n}. \tag{6-32}$$

For a long line, where 2βns > 3:

$$B_{0n} = B_0 + \ln \sqrt{4\beta s}. \tag{6-33}$$

Since the quantity $\sqrt{4\beta s}$ is less than one, $B_{0n} < B_0$.

Figure 6-20 shows the dependence of the near-end cross-talk attenuation on the length of the cable line (the number of shipping-length cable sections). The figure makes it clear that B_{0n} decreases as the line becomes longer, but at some particular length becomes stable and remains equal to

$$B_0 + \ln \sqrt{4\beta s}.$$

The formula for computing the cross-talk attenuation at the near end of a long line (6-33) can be represented in another form. Since $B_0 = \ln |2/\omega Z K_0|$, then

$$B_0 = \ln \left|\frac{2}{\omega Z K_0}\right| + \ln \sqrt{4\beta s} = \ln \left|\frac{4\sqrt{\beta s}}{\omega Z K_0}\right|. \qquad (6\text{-}34)$$

With respect to intersection at the far end, the interference-current element I^k, leaving any shipping-length section of the cable, sees practically the same attenuation, and may be expressed as:

$$I^k_{2i} = I_1 e^{-\beta n s} e^{-B_l},$$

where I_1 is the current in the input of the disturbing circuit;

βns is the attenuation of the entire cable circuit itself;

B_1 is the cross-talk attenuation between the first and second circuits per standard (shipping) section of the cable.

The resultant interference current flowing at the far end of the second circuit and equal to the geometric sum of the elementary currents is:

$$I_{2l}=\sqrt{\sum_{n=1}^{n}(I_{2l}^{k})^2}=\sqrt{(I_1 e^{-\beta ns} e^{-B_l})^2 n}=I_1 e^{-\beta ns} e^{-B_l}\sqrt{n},$$

where n is the total number of sections (shipping lengths of cable).

The cross-talk attenuation at the far end of the entire cable circuit will thus be:

$$B_{ln}=\ln\frac{I_1}{I_{2l}}=B_l+\beta ns-\ln\sqrt{n}, \qquad (6\text{-}35)$$

here $\ln\sqrt{n}$ acts to decrease the cross-talk attenuation as the length of the line increases.

Up to a certain length of line, $\ln\sqrt{n}$ has little effect, so that at first B_{ln} drops; then, owing to the increasing attenuation of the line itself, it sharply rises.

When the line is not too long, the quantities B_{0n} and B_{ln} are nearly equal, while for repeater sections of cable circuits (βns = 4 to 6 nepers) the cross-talk attenuation is always greater at the far end than at the near end. This is confirmed by Table 6-10, which gives the results of measurements of cross-talk attenuation for 92-km cable lines consisting of 368 shipping lengths of 250 m each.

Thus, cable communications circuits should be used with a system under which the quality of communication is

determined by the interaction at the far end of the cable, i. e., either a four-wire--two-cable or electrically two-wire--one-cable system.

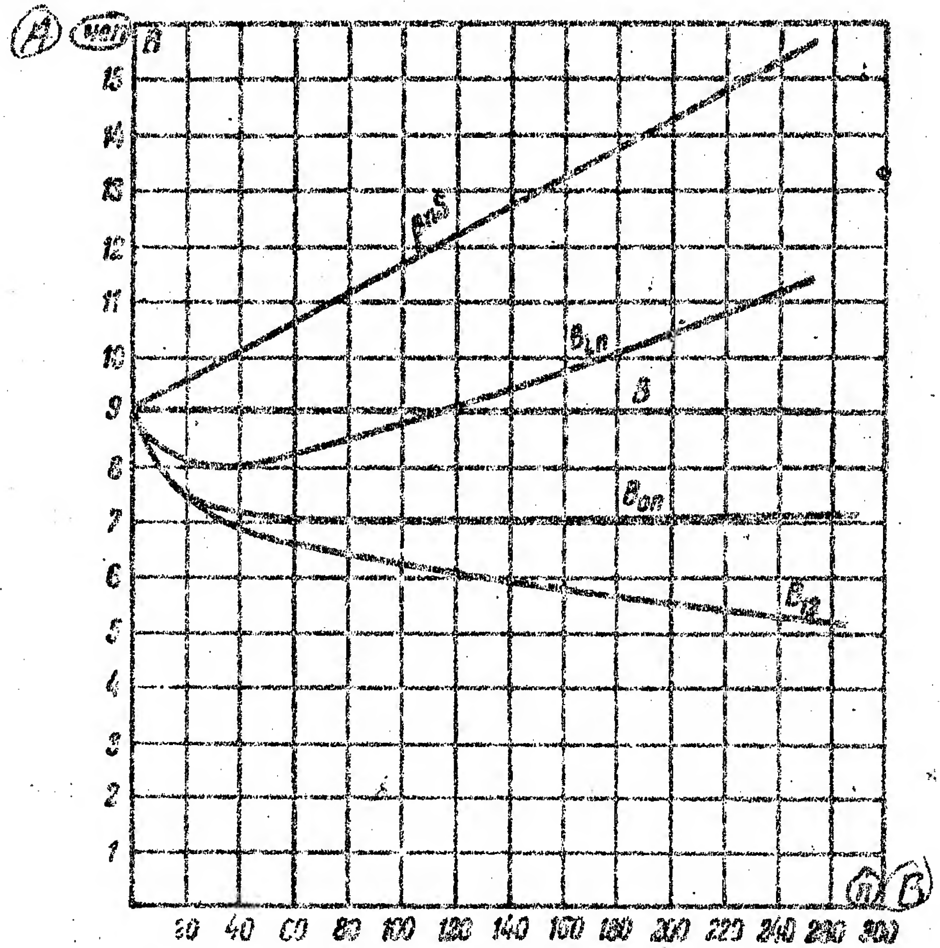


Fig. 6-20. Cross-talk attenuation in a cable line. A) nepers; B) n.

Knowing that $B_1 = \ln |2/\omega Z K_1|$, the quantity B_{1n} may be represented as

$$B_{1n} = \ln \left| \frac{2}{\omega Z K_1 \sqrt{n}} \right| + \beta ns. \qquad (6-36)$$

Turning to the quantity B_{12} used in wire-communications technology (the difference in the levels of the received and interference currents), using the terminology of the International Consultative Committee, the interference resistance of circuits can be expressed in the following manner:

$$B_{12} = B_{ln} - \beta ns = B_l - \ln \sqrt{n} = \ln \left| \frac{2}{\omega Z K_l \sqrt{n}} \right|. \qquad (6\text{-}37)$$

The interference resistance of a shipping-length cable, where the attenuation of the line itself may be disregarded, equals the cross-talk attenuation at the far end:

$$B_{12} = B_l = \ln \left| \frac{2}{\omega Z K_l} \right|. \qquad (6\text{-}38)$$

As the length of the line increase, the resistance of the cable circuit decreases according to the $\ln \sqrt{n}$ law. As the frequency increases, the quantity B_{12} decreases, following an approximately logarithmic law (Table 6-10).

Table 6-10. Results of measurement of the cross-talk attenuation of a 92-km long cable (in nepers). A) Frequency, cps; B) Near-end cross-talk attenuation; C) Far-end cross-talk attenuation; D) Circuit resistance to noise; E) Attenuation of the line itself.

Частота, кгц (A)	1	2	3	3,5	4	4,7	6,3	7,5	10,2	11,5	13	15	20
(B) Переходное затухание на ближнем конце	>10	9,6	8,2	8,2	8,7	9,1	9,3	8,2	8,3	7,5	8,2	8,2	7,2
(C) Переходное затухание на дальнем конце	>10	>10	>10	>10	>10	>10	>10	>10	>10	>10	>10	10	10
(D) Защищенность цепей	>10	>10	>10	10	10	9,9	9,8	9,5	—	—	—	—	—
(E) Собственное затухание . .	2,1	2,7	3,2	3,6	4,0	4,3	5,3	—	—	—	—	—	—

6-8. CCI STANDARDS FOR INTERFERENCE REJECTION ON LONG-LINE CABLES

The Internation Consultative Committee (CCI) has standardized the following quantities for capacitive and inductive coupling and cross-talk attenuation on low-frequency (up to 3,000 cps) and high-frequency (up to 60,000 cps and above) cable links.

1. The capacitive-coupling coefficient and the capacitive unbalance for a shipping-length section of cable

230 m long, measured at a frequency of 800 cps, should not exceed the values shown in Table 6-11.

Table 6-11. Values of capacitive coupling and capacitive unbalance coefficients for low-frequency cables. A) coefficient; B) quad relationship; C) circuits; D) values of capacitive coupling and unbalance; E) mean, uuf; F) maximum, uuf; G) within the quads; H) physical/physical; I) Phantom/physical; J) physical/ground; K) phantom/ground; L) between quads; M) physical/physical; N) phantom/physical; O) phantom/phantom.

(A) Коэффициент	(B) Какие четверки	(C) Наименование цепей	(D) Значения емкостной связи и асимметрии	
			(E) Среднее, пф	(F) Максимальное, пф
k_1	(G) Внутри четверки	(H) Основная/основная	40	150
$k_2 - k_3$		(I) Фантомная/основная	75	375
$e_1 - e_2$		(J) Основная/земля	150	600
e_3		(K) Фантомная/земля	300	1 200
$k_9 - k_{12}$	(L) Между четверками	(M) Основная/основная	60	225
$k_5 - k_8$		(N) Фантомная/основная	60	225
k_4		(O) Фантомная/фантомная	60	225

2. For high-frequency multiplexed cables, the capacitive coupling and capacitive unbalance coefficients, measured at 800 cps, on a shipping-length cable 230 m long,

should not exceed the values shown in Table 6-12.

Table 6-12. Values of Capacitive coupling coefficients for high-frequency cables. A) Coefficient; B) Between circuits; C) Values of capacitive coupling and unbalance, $\mu\mu f$; D) Mean; E) Maximum; E') Between the circuits of a single quad; F) Between quads of adjacent quads with the same lay; G) Between circuits of nonadjacent quads with the same lay; H) Between circuits of quads of adjacent layers; I) between a circuit and ground.

(A) Коэффициент	(B) Между цепями	(C) Значения емкостной связи и асимметрии, пф (D) Среднее	(E) Максимальное
k_1	(E') Между цепями одной четверки . . .	33	125
$k_9 - k_{12}$	(F) Между цепями смежных четверок того же повива	10	60
$k_9 - k_{12}$	(G) Между цепями несмежных четверок того же повива	—	20
$k_9 - k_{12}$	(H) Между цепями четверок смежных повивов	10	60
$e_1 - e_2$	(I) Между цепью и землей	100	400

If the shipping length of the cable $l \neq 230$ m, then the capacitive coupling values should not exceed the values determined by means of the following corrections:

a) The mean values k_1, k_4, k_{5-8}, k_{9-12} are multiplied by $\sqrt{l/230}$.

b) The maximum values k_1, k_4, k_{5-8}, k_{9-12}, and also the mean and maximum values of k_2, k_3, e_1, e_2, e_3 are multiplied by $l/230$.

3. The inductive-coupling coefficients in high-frequency cable links, measured at 5,000 cps on a shipping-length (230 m) line, should not exceed the values shown in Table 6-13.

4. In an assembled cable, the values of the capacitive coupling and the capacitive unbalance, measured at 800 cps over a distance equal to the coil-loading spacing, should not exceed the values shown in Table 6-14.

5. For shipping-length sections, the resistive unbalance of conductors should not exceed 1%.

In assembled cables, the resistive unbalance of the conductors within the coil-loading spacing should not exceed 0.25%.

Table 6-13. Values of Inductive-coupling coefficients. A) Coefficient; B) Between circuits; C) Values of inductive coupling, millimicrohenrys; D) Mean; E) Maximum; F) Between circuits of a single quad; G) Between circuits of adjacent quads with the same lay; H) Between circuits of non-adjacent quads with the same lay; I) Between circuits of quads of adjacent layers.

(A) Коэффициент	(B) Между цепями	(C) Значение индуктивной связи, нгн	
		(D) Среднее	(E) Максимальное
m_1	(F) Между цепями одной четверки . . .	150	600
$m_9 - m_{12}$	(G) Между цепями смежных четверок того же повива	100	400
$m_9 - m_{12}$	(H) Между цепями несмежных четверок того же повива	50	350
$m_9 - m_{12}$	(I) Между цепями четверок смежных повивов	100	600

Table 6-14. Values of the capacitive coupling and capacitive unbalance coefficients for assembled cables of high- and low-frequency links. A) Coefficient; B) Between circuits; C) Values of capacitive coupling and unbalance, $\mu\mu f$; D) Mean; E) Maximum; F) Physical/physical; G) Phantom/physical; H) Physical/ground; I) Phantom/ground; J) Physical/physical; K) Phantom/physical; L) Phantom/phantom.

A Коэффициенты	B Между цепями	C Значения емкостной связи и асимметрии, пф	
		D Среднее	E Максимальное
k_1	F Основная/основная	10	20
$k_2 - k_3$	G Фантомная/основная	10	20
$e_1 - e_2$	H Основная/земля	100	300
e_3	I Фантомная/земля	200	500
$k_9 - k_{12}$	J Основная/основная	—	40
$k_5 - k_8$	K Фантомная/основная	—	60
k_4	L Фантомная/фантомная	—	80

The percentage unbalance can be computed from the following formula:

$$\frac{2(R_a - R_b)}{R_a + R_b} 100\%, \qquad (6\text{-}39)$$

where R_a and R_b are the respective resistances of the circuit conductors.

6. The deviation of the effective capacitance from the mean value over a shipping-length section should not exceed 4% (average) and 12.5% (maximum).

In an assembled cable, the deviation of the effective capacitance over the coil-loading spacing should not exceed 0.8% (average) and 2% (maximum).

The percent capacitive deviation is computed according to the formula

$$\frac{C_a - C_n}{C_n} 100\% \qquad (6\text{-}40)$$

where C_a is the actual capacitance of the circuit;

C_n is the nominal (average for all circuits) capacitance.

7. The resistance to mutual interference (the difference in level between the useful signal and the interference) over the entire range of frequencies transmitted, for a repeater section, should not fall below the values shown in Table 6-15.

Table 6-15. Standards for noise-resistance B_{12}, in nepers, for a repeater section of trunk cable. A) Type of link; B) Resistance B_{12}, nepers; C) Two-wire circuit; D) Four-wire circuit; E) Radio broadcast circuit.

(A) Вид связи	Защищенность B_{12}, неп (B)
(C) Двухпроводная цепь . . .	7
(D) Четырехпроводная цепь . .	7,5
(E) Радиовещательная цепь . .	9,5

The cross-talk attenuation at the far and near ends, B_{1n} and B_{0n}, separately should not be less than the sum of the interference resistance B_{12} and the circuit attenuation βl:

$$B_{1n} = B_{0n} = B_{12} + \beta l. \tag{6-41}$$

If it is necessary to define the standard interference resistance B^*_{12} for shipping or other lengths of cable l_x, the conversion is carried out on the basis of the law for geometric addition of interaction currents from the individual cable sections:

$$B^*_{12} = B_{12} + \ln \sqrt{\frac{l_{y.y}}{l_x}}, \tag{6-42}$$

where B^*_{12} is the resistance to interference for the defined section of cable;

B_{12} is the resistance to interference for a repeater section (see Table 6-15);

$l_{y,y}$ is the length of the repeater section;

l_x is the length of the defined cable section.

6-9. THE CHARACTERISTIC COUPLING RATIOS AND THEIR EFFECT ON THE CABLE CROSS-TALK ATTENUATION

It has been established that in cables there is a definite relationship between the inductive and capacitive couplings. It depends upon the type of twist, and is called the characteristic ratio:

$x = m_1/k_1$ [henrys/farad].

Any deviation of x from the standard values indicates a defect in the design or construction of the cable lay-up. In particular, in spiral-lay cable, the characteristic ratio has the following value.

Table 6-16. A) Frequency, cps; B) [Henrys/farad]; C) Cable with styroflex insulation; D) Cable with paper insulation.

Частота, кгц (A)	$x = \frac{m_1}{k_1}\left(\frac{гн}{ф}\right)$ (B)	
	Кабель со стирофлексной изоляцией (C)	Кабель с бумажной изоляцией (D)
0,8	6 000—8 000	7 000—9 000
13,5	6 000—7 000	6 000—8 000
60,0	5 000—6 000	6 000—7 000

As the frequency increases, the characteristic ratio drops, owing to the decreasing value of the inductive coupling.

In DP twist cables, the characteristic ratio reaches

10,000 to 15,000 henrys/farad.

Let us consider the dependence of the cross-talk attenuation on the ratio of the electrical and inductive coupling coefficients; for clarity, we will determine not the quantity $x = m_1/k_1$, but the components of the electromagnetic coupling coefficients

$$K_0 = K_c + K_m,$$
$$K_i = K_c - K_m.$$

Without considering the in-phase components of coupling, the ratio of the effective inductive coupling K_m and the effective capacitive coupling K_c is expressed as

$$y = \frac{K_m}{K_c} = \frac{\frac{m_1}{Z^2}}{\frac{k_1}{4}} = \frac{m_1}{k_1} \cdot \frac{4}{Z^2} = x\,\frac{4}{Z^2}. \tag{6-43}$$

Keeping in mind that $B_0 = \ln |2/\omega Z K_0|$, while the coefficient $K_0 = (k_1/4) + (m/Z^2)$, we obtain

$$e^{-B_0} = \left|\frac{\omega Z}{2}\left(\frac{k_1}{4} + \frac{m_1}{Z^2}\right)\right|.$$

Where there is no inductive coupling, the cross-talk attenuation is determined by the capacitive coupling alone, and is expressed by the formula

$$e^{-B0k} = \left|\frac{\omega Z}{2} \cdot \frac{k_1}{4}\right| = \left|\frac{\omega Z k_1}{8}\right|.$$

Where there is no capacitive coupling, the cross-talk attenuation takes the form

$$e^{-B_{0m}} = \left|\frac{\omega Z}{2} \cdot \frac{m_1}{Z^2}\right| = \left|\frac{\omega m_1}{2Z}\right|.$$

Dividing the quantity e^{-B_0} by $e^{-B_{0k}}$ and $e^{-B_{0m}}$ in turn, we have

$$\frac{e^{-B_0}}{e^{-B_{0k}}} = e^{B_{0k}-B_0} = 1 + \frac{m_1}{k_1} \cdot \frac{4}{Z^2} = 1 + y,$$

$$\frac{e^{-B_0}}{e^{-B_{0m}}} = e^{B_{0m}-B_0} = 1 + \frac{k_1}{m_1} \cdot \frac{Z^2}{4} = 1 + \frac{1}{y}$$

or, after taking logs, we obtain at the near end

$$B_{0k} - B_0 = \ln(1+y); \tag{6-44}$$

$$B_{0m} - B_0 = \ln\left(1 + \frac{1}{y}\right) \tag{6-45}$$

and at the far end of the cable

$$B_{lk} - B_l = \ln(1-y); \tag{6-46}$$

$$B_{lm} - B_0 = \ln\left(\frac{1}{y} - 1\right) \tag{6-47}$$

where the quantities B_{0k} and B_{1k} are the cross-talk attenuations due to capacitive coupling alone (where there is no inductive coupling);

B_{0m} and B_{1m} are the cross-talk attenuations for inductive coupling alone;

B_0 and B_1 are the cross-talk attenuations for the case where both types of coupling act together.

The quantities $B_{0k} - B_0$ and $B_{1k} - B_1$ indicate that the cross-talk attenuation drops somewhat at the near and far ends of the cable owing to the inductive coupling.

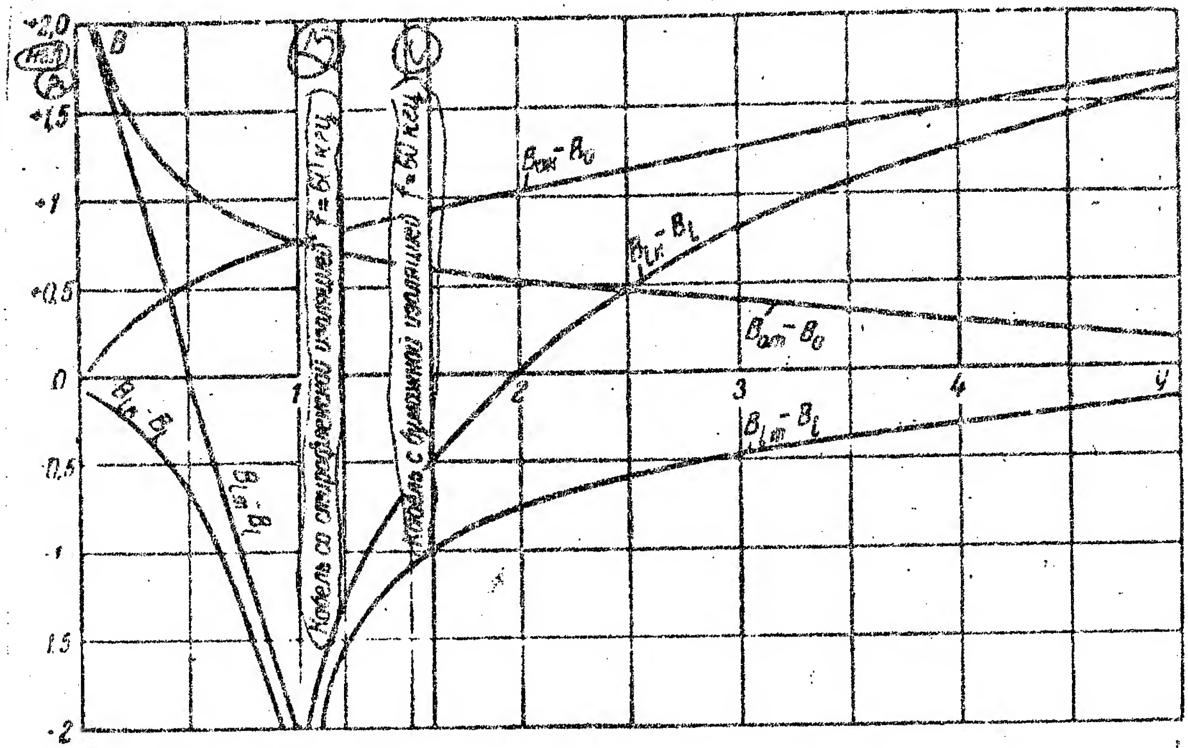


Fig. 6-21. The cross-talk attenuation as a function of the coupling ratio. A) nepers; B) Cable with styroflex insulation, f = 60 kc; C) Cable with paper insulation, 60 kc

$B_{0m} - B_0$ and $B_{1m} - B_1$ characterize the decrease in cross-talk attenuation owing to capacitive coupling.

Figure 6-21 gives the dependence of these quantities

on the coupling ratio y (6-43) in cables. The figure also gives practical values of y for cables with paper-cord and styroflex-cord insulation.

From the data given, it follows that:

1. For interaction at the near end of a cable, the cross-talk attenuation drops as the capacitive or inductive coupling increases.

2. For interaction at the far end of the cable, increasing y toward one gives a positive effect. At $y = 1$, the quantities $B_{1m} - B_1$ and $B_{1k} - B_1$ approach infinity. In this case, the capacitive and inductive coupling cancel, and the interaction of the circuits approaches a minimum.

For interaction at the far end, the cross-talk attenuation is greater in the presence of both couplings, B_1, than for B_{1k} or B_{1m} alone.

At the values $y = 0.5$ and $y = 2$, the quantities $B_{1m} - B_1$ and $B_{1k} - B_1$ are equal to zero, respectively, i.e., in these cases, the cross-talk attenuation with both couplings acting equals the cross-talk attenuation for the coupling B_{1m} or B_{1k} acting alone.

3. The actual values of y for paper cord and styroflex insulation are 1.2 and 1.6 respectively. In cable with styroflex cord insulation, y is close to one, which gives the best cross-talk attenuation at the far end.

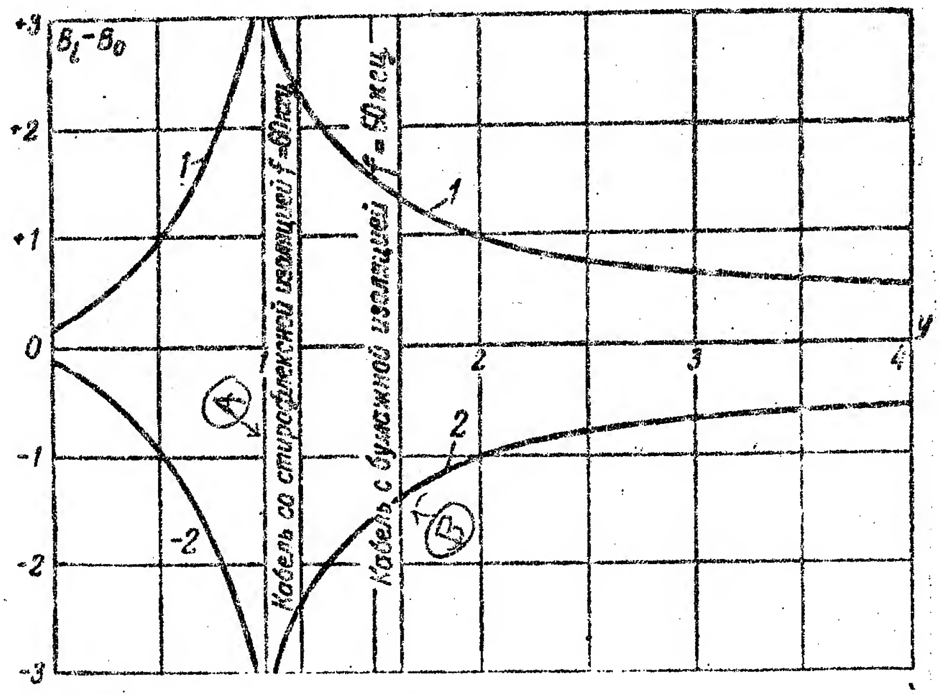


Fig. 6-22. The variation of $B_1 - B_0$ as a function of the coupling ratio. 1) Where m and k have the same sign; 2) where m and k have differing signs. A) Cable with styroflex insulation, f = 60 kc; B) Cable with paper insulation, f = 60 kc.

The difference between the cross-talk attenuation of cables with styroflex and paper insulation is: a) 0.1 to 0.4 nepers at the near end; b) 05 to 0.75 at the far end.

4. From the point of view of decreasing the interaction at the near end, it is desirable to have y equal to 0, for $B_{0k} - B_0$, and to ∞ for $B_{0m} - B_0$. Mini-

mum interaction at the far end occurs when $y = 1$.

Let us consider the relationship between B_0 and B_1 and decide upon the most advantageous value of coupling at the far and near ends of the cable

$$B_l - B_0 = \ln\left|\frac{2}{\omega Z K_l}\right| - \ln\left|\frac{2}{\omega Z K_0}\right| = \ln\left|\frac{K_0}{K_l}\right|$$

or

$$B_l - B_0 = \ln\frac{\frac{k_1}{4} + \frac{m_1}{Z^2}}{\frac{k_1}{4} - \frac{m_2}{Z^2}} = \ln\frac{1+y}{1-y}. \qquad (6\text{-}48)$$

Figure 6-22 gives the dependence of the quantity $B_1 - B_0$ on y for both like and unlike signs of the capacitive and inductive coupling coefficients. It is clear from the figure that when the signs are the same, the cross-talk attenuation is greater at the far end than at the near end, i.e., $B_1 - B_0 > 1$. When $y = 1$, the quantity $B_1 - B_0$ is theoretically equal to infinity at the far end, showing that the capacitive and magnetic coupling cancel when the coefficients have the same sign.

When m_1 and k_1 have unlike signs, the curves of $B_1 - B_0$ are mirror images of the curves of $B_1 - B_0$ when the signs are the same.

Since in the great majority of cases the capacitive and inductive coupling coefficients have the same sign, the quantity B_1 is greater than B_0, as a rule.

Therefore, it is necessary to give preference to that system of long-distance cable communication for which the quality of transmission is determined by the interaction at the far end.

6-10. MEASURING THE ELECTROMAGNETIC COUPLING COEFFICIENTS DURING THE MANUFACTURING PROCESS

Let us consider the effect of various production operations on the coupling coefficients in cables, and find the degree to which these coefficients change during the manufacturing process.

Table 6-17 gives data for measurements of the capacitive coupling coefficients k_1 for cables with styroflex-cord insulation (4 X 4) during the manufacturing process (up to lay-up, after lay up, after the application of the lead, and after the cable has been armored). It also gives the values of the effective capacitance of cables.

Figures 6-23 and 6-24 show static diagrams of the capacitive coupling k_1 and the inductive coupling m_1 in type 32 X 2 insulated cables which, for these measurements, were 30 shipping lengths long.

The values of the coupling coefficients are plotted along the axis of abscissas, and the percentage of quads whose coupling coefficients fall below the given value along the axis of ordinates.

The diagrams indicate that the lead covering of the cables stabilizes and decreases the coupling coefficients. Thus, where before lead-coating the percentage of quads having $k_1 \leq 10$ $\mu\mu f$ was 10%, after lead-coating, it increases to 48%.

The technical requirements with respect to the values of k_1 and m_1 were met by 95-97% of paper-cord insulated cables.

It is also clear from Table 6-17 that the lay of the quads in the cable also decreases the capacitive unbalance in the cable.

After the lead is applied, the effective capacitance rises 0.4 to 1.5 millimicrofarads per kilometer.

The distribution of the capacitive coupling coefficients (k_1, k_{9-12}, e_1), based on the of 200 measurements on shipping-length sections of styroflex cord insulated cable, is shown in Figs. 6-25, 6-26, and 6-27:

Table 6-17. Date on the measurement of the effective capacitance and capacitive coupling coefficients in quads in the cable manufacturing process. A) Drum number; B) Quad number: C) Pair Number; D) lay-up; E) After lay-up; F) After lead-coating; G) After armoring; H) Effective capacitance, millimicrofarads; I) Per shipping length; J) Per km of finished cable; K) Intraquad capacitive coupling coefficient, $\mu\mu f$; L) Before lay-up; M) After lay-up; N) After lead-coating; O) After armoring.

Table 6-17

№ барабана (A)	№ четверок (B)	№ пар (C)	Рабочая емкость, пф (H): На строительной длине (I): До скрутки (D)	После скрутки (E)	После освинцевания (F)	После бронирования (G)	На 1 км в готовом кабеле (J)	Коэффициенты емкостной связи внутри четверки, пф (K): До скрутки (L)	После скрутки (M)	После освинцевания (N)	После бронирования (O)
1	2	3	4	5	6	7	8	9	10	11	12
	246	1	6,9	7,0	7,15	7,02	27,6	3	1	—2	—0,5
		2	6,95	7,05	7,15	7,05	27,75				
74	251	1	6,8	7,02	7,15	7,05	27,75	6	—4	—10	—8,1
		2	6,86	7,07	7,07	7,06	27,8				
	304	1	6,5	6,95	7,12	7,01	27,6	6	—18	—26	—18,5
		2	6,5	7,0	7,18	7,05	27,75				
	269	1	7,0	6,9	7,26	7,16	28,2	30	—18	—10	27,2
		2	7,1	6,95	7,3	7,2	28,3				
85	284	1	6,9	6,97	7,35	7,36	28,95	25	27	6	22,0
		2	6,7	7,05	7,4	7,4	29,1				
	291	1	6,75	6,88	7,3	7,15	28,15	—26	—11	—10	—11
		2	6,9	7,01	7,37	7,35	28,9				
	385	1	6,55	7,1	7,35	7,35	28,9	—30	—10	—3,2	—25
		2	6,36	6,94	7,2	7,2	28,2				
106	386	1	6,55	7,0	7,25	7,24	28,45	—36	1	6	—15
		2	6,6	7,02	7,25	7,26	28,55				
	300	1	6,45	6,96	7,2	7,2	28,2	39	—30	7,5	24
		2	6,51	7,04	7,24	7,25	28,5				
	389	1	6,65	6,98	7,2	7,04	27,7	23	—4	—3,8	—1,5
		2	6,75	7,01	7,22	7,04	28,7				
	382	1	6,65	6,97	7,18	7,0	27,55	—38	—7	—12	—28
		2	6,67	7,01	7,22	7,0	27,55				
	365	1	6,33	6,95	7,15	7,0	27,55	27	8	24,5	39,5
		2	6,4	6,99	7,21	7,07	27,8				

It follows from the figures that for the majority of cables, $k_1 = 4\ \mu\mu f$, $k_{9-12} = 7.8\ \mu\mu f$; and $e_1 = 10\ \mu\mu f$.

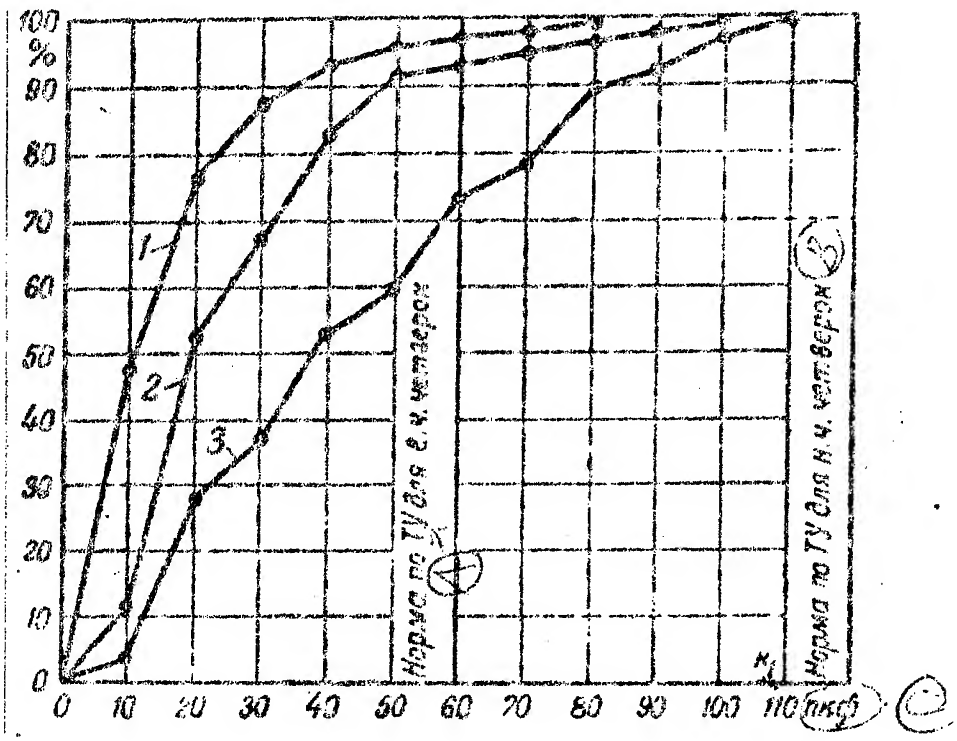


Fig. 6-23. Static curves for the coefficient of coupling. 1) After application of lead; 2) after balancing; 3) before balancing; A) Technical specification standard for high-frequency quad; B) Technical specification standard for low-frequency quad; C) Micromicrofarad [the Russian here has an abbreviation that would yield 'micromicrokilofarad' if taken literally; it would seem to be a typographical error].

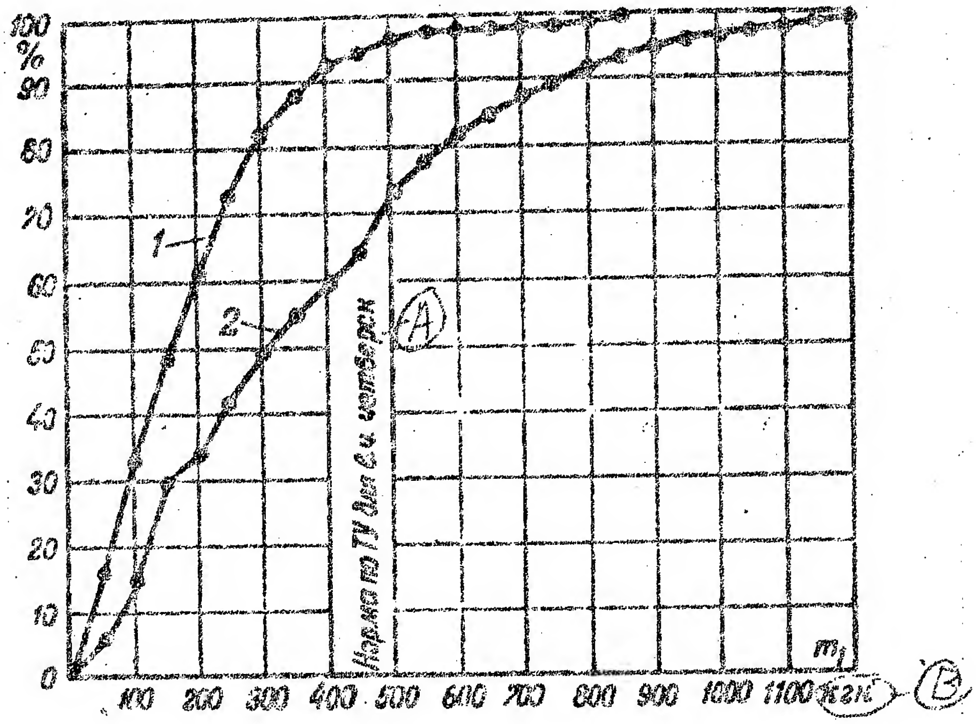


Fig. 6-24. Static curves for coupling coefficient. 1) After lead has been applied; 2) before balancing; A) Technical-specification standard for high-frequency quad; B) kilohenrys [another apparent error; most likely should read 'millimicrohenrys'].

The maximum value of k_1 and k_{9-12} does not exceed 15 to 16 μμf, i. e., they do not exceed the standard values for these cables of $k_1 \leqslant 25$ μμf and $k_{9-12} \leqslant 20$ μμf.

Repeated winding of styroflex-insulated cable on standard drums at a rate of 15 to 25 meters/minute had a negligible effect its electrical properties.

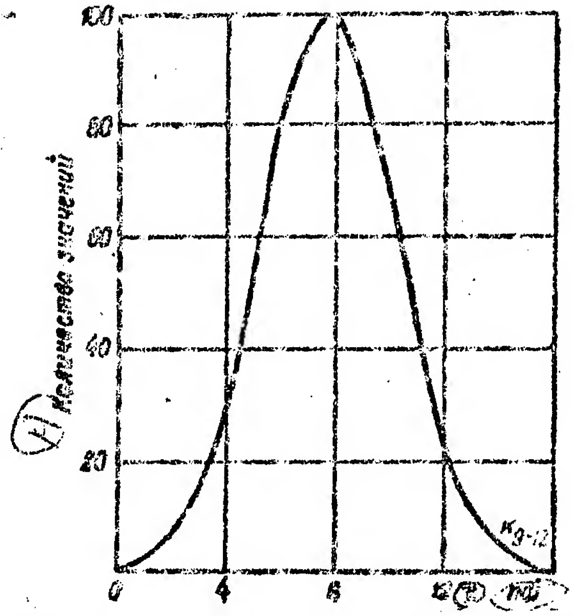


Fig. 6-25. Distribution of the average values of the capacitive coupling coefficients k_{9-12} in 200 shipping-length sections. A) Number of values; B) $\mu\mu f$.

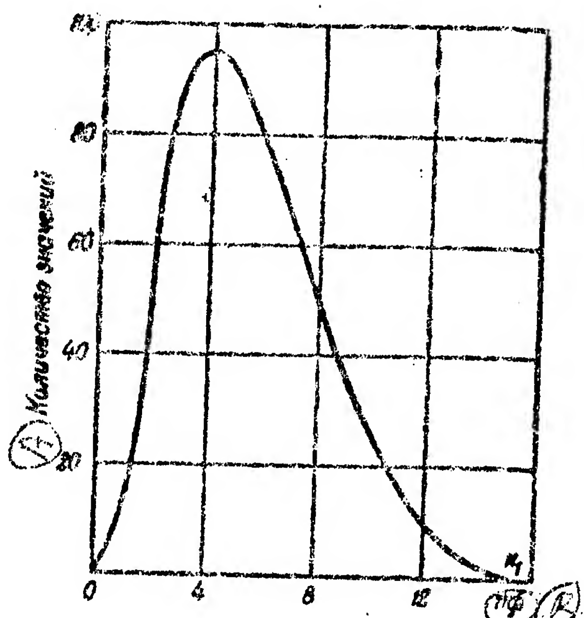


Fig. 6-26. Distribution of average values of capacitive-coupling coefficient k_1 in 200 shipping-length sections. A) Number of values; B) $\mu\mu f$.

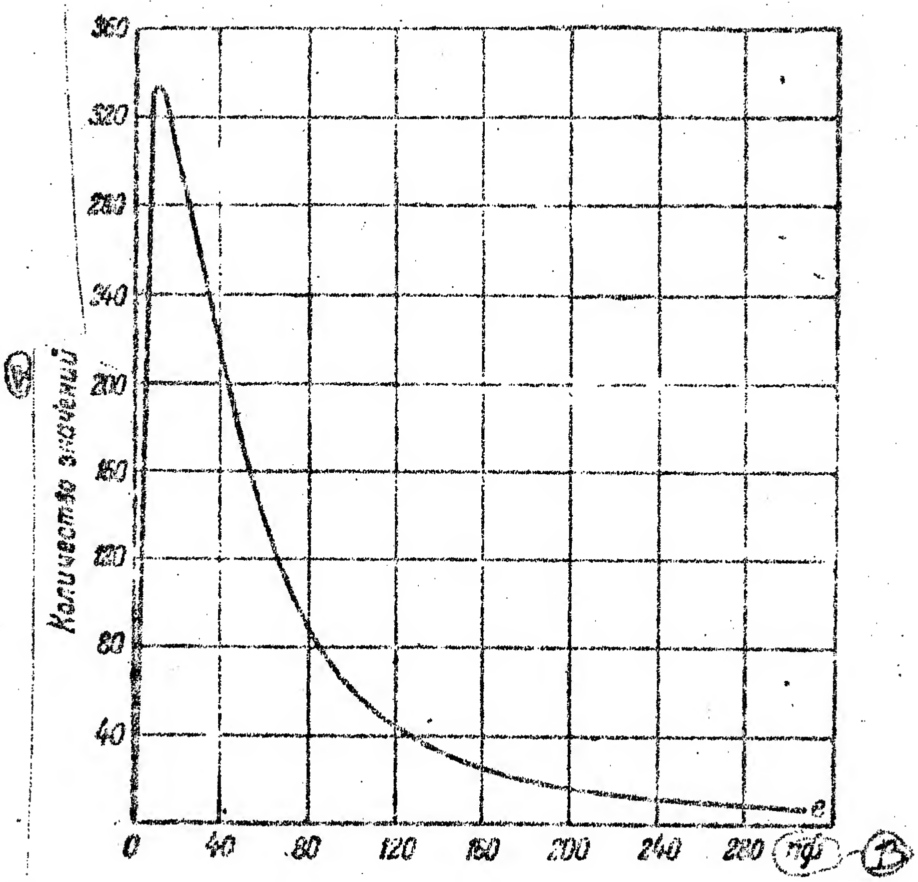


Fig. 6-27. Distribution of average values of ground capacitive coupling coefficients in 200 shipping-length sections. A) Number of values; B) μμf.

6-11. INDIRECT INTERACTION BETWEEN CIRCUITS

Until now, we have considered the process of interaction between circuits in somewhat idealized form, since we have dealt with uniform lines with matched loads at the ends.

Under actual conditions, there are internal nonuni-

formities in the cable, individual shipping length sections differ somewhat in their properties, and in addition, the impedances of the loads at the ends may not match the wave (characteristic) impedance of the cable ($Z_{np} \neq Z$). This would cause wave reflections at the points of electrical mismatch.

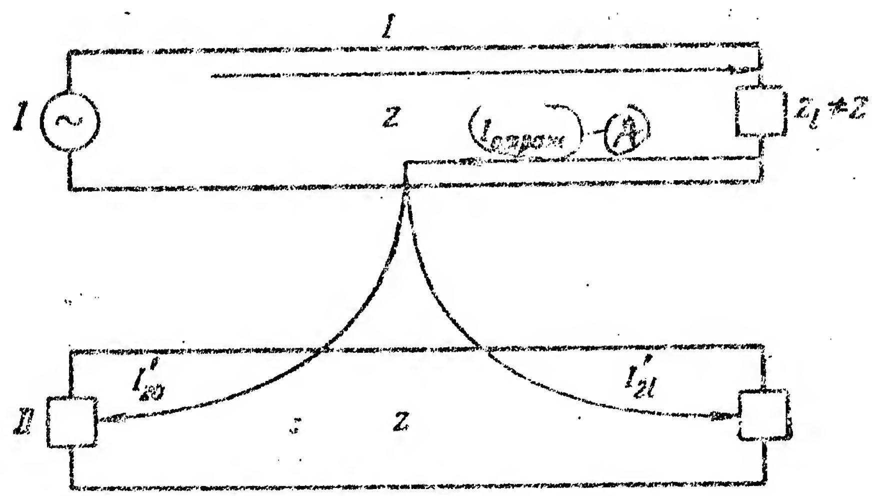


Fig. 6-28. Interaction between circuits owing to mismatch between the load impedance Z_{np} and the wave impedance of the cable, Z. A) $I_{reflect}$.

It has already been shown that the presence of reflections involves a change(generally an increase) in the attenuation of the cable itself, and is associated with distortion of the signals transmitted over the cable.

Cable nonuniformities and mismatches with equipment

lead to the appearance of additional interaction between the circuits and a decrease of the cross-talk attenuation. In addition to interaction due to nonuniformities, it is also possible for energy to be transferred through an adjacent circuit. Such energy transfer is distinguished from direct interaction of I and II circuits, and is called interaction through a third circuit (I - III and III - II).

All of these interactions (due to line nonuniformities, load mismatches, and through a third circuit) are called indirect effects to distinguish them from the direct effects which we have considered previously.

Figure 6-28 shows how additional interaction arises between circuits where there is a load mismatch (Z_{np}) with the wave impedance Z of the cable ($Z_{np} \neq Z$).

The electromagnetic energy appearing at the end of circuit I is only partially transmitted to the receiver, owing to the load mismatch; part of it is reflected back to the line input.

The reflected energy, turned back along circuit I on account of the capacitive and inductive coupling between the circuits, is in part transferred to circuit II, and appears at the far and near ends in the form of a noise current.

Thus, besides the current resulting from direct

effects, there is additional interaction owing to the reflection of energy when there is a load mismatch.

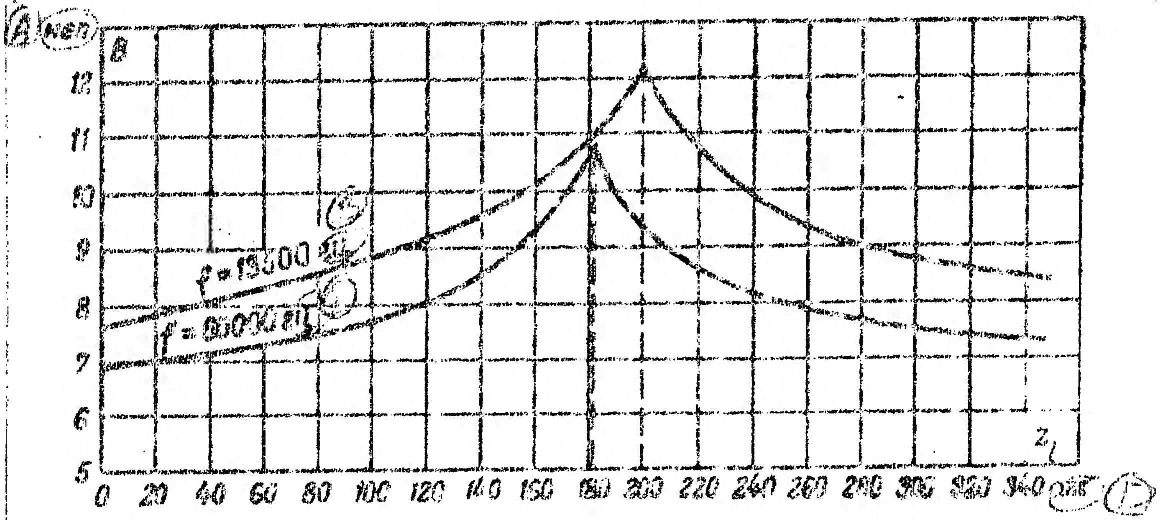


Fig. 6-29. Cross-talk attenuation as a function of load impedance. A) Nepers; B) cps; C) cps; D) ohms.

A similar process occurs when the load is mismatched to the cable Z at the input and output ends of circuit II.

The larger the value of the reflection coefficient p, the more interaction between the circuits:

$$p = \frac{Z_{np} - Z}{Z_{np} + Z}.$$

The decrease in cross-talk attenuation occasioned by reflections is characterized by the mismatch attenuation parameter B_{ref}:

$$B_{ref} = \ln\left|\frac{1}{p}\right| = \ln\left|\frac{Z_{np} + Z}{Z_{np} - Z}\right|. \qquad (6\text{-}49)$$

According to the existing standards for high-frequency circuits, the reflection coefficient should not exceed $1.25/\sqrt[4]{f}$, where f is the frequency in kc.

Figure 6-29 shows the results of measurements of the cross-talk impedance for shipping-length sections of cable with paper-cord insulation for various load impedances at the ends of the disturbing and disturbed circuits of the cable. It is clear from Fig. 6-29 that the largest value of cross-talk attenuation is obtained when the loads are matched (at f = 60,000 cps, z for the cable is 182 ohms). Both lowering and raising the load impedance lowers the cross-talk attenuation. Thus, if at 60,000 cps with a matched load, B = 10.8 nepers, then at Z_{np} = 100 ohms, B will have a value of only 7.7 nepers.

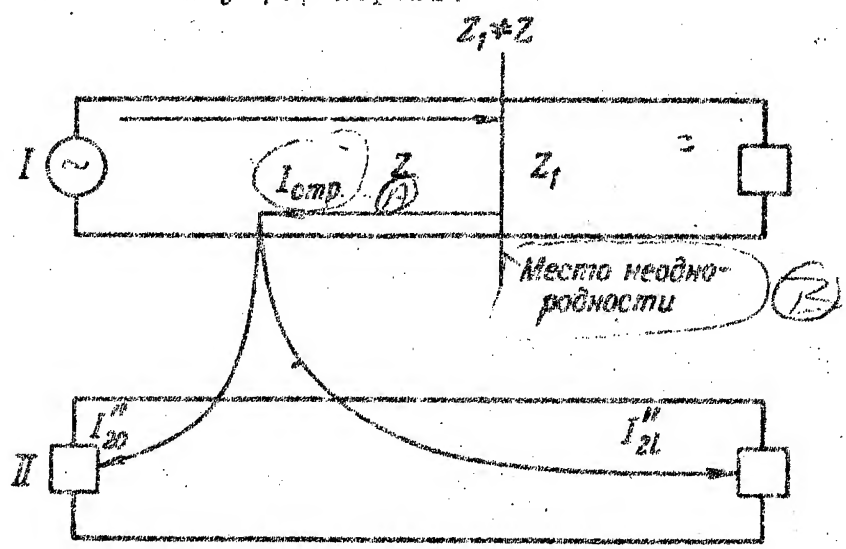


Fig. 6-30. Interaction between circuits owing to internal nonuniformities in the cable, $Z_1 \neq Z$. A) I_{ref}; B) Point of nonuniformity.

In the lower frequency region, there is a still sharper change in the value of the cross-talk attenuation owing to load mismatches.

The additional circuit interaction owing to cable nonuniformities arises in the manner considered above.

Here, just as in the case of mismatching of the load impedance to the wave impedance of the cable circuit ($Z_1 \neq Z$), reflected waves appear which increase the interaction currents of the circuits. The less uniform the cable is along its length, the greater the local energy reflections, and the greater the intercircuit interference. The value of cross-talk attenuation is correspondingly decreased. Additional interaction at the far and near ends of the second circuit owing to internal nonuniformities in the first circuit ($Z_1 = Z$) is illustrated in Fig. 6-30.

Let us consider the phenomenon of interaction of circuits I and II due to the presence of circuit III.

By circuit III, we mean one loop or another lying in the same cable as circuits I and II.

As can be seen from Fig. 6-31, the disturbing circuit I induces noise currents at the near and far ends of circuit III. In turn, circuit III becomes a source of interference, and sets up noise currents in circuit II.

Thus, the indirect effect of circuit I on circuit II takes place in two stages (I - III and then III - II).

The interaction by way of circuit III shows up at both the near and far ends of circuit II; it has been theoretically and experimentally established, however, that most of the noise arrives at the far end of circuit II by way of the far end of circuit III.

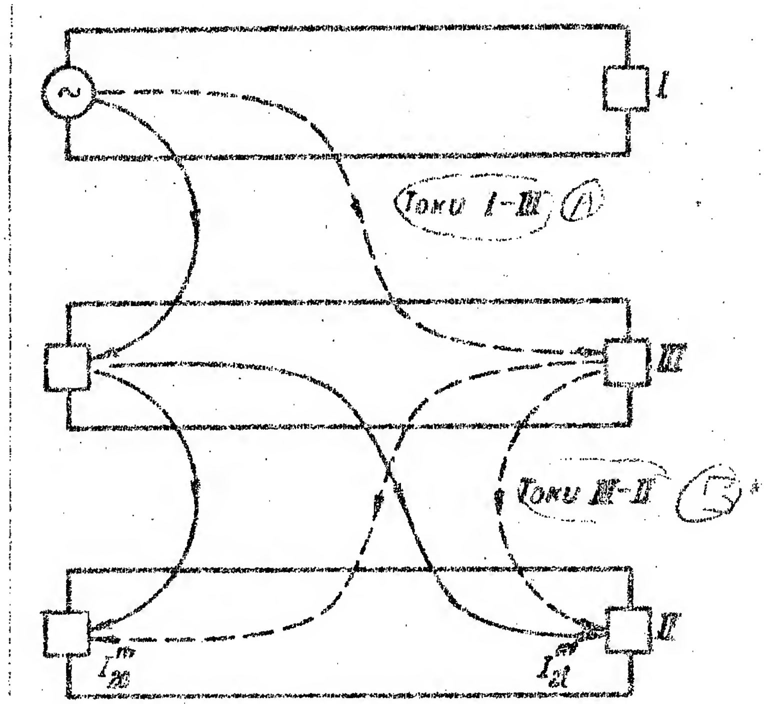


Fig. 6-31. Interaction of circuits I and II by way of a third circuit III. A) I - III currents; B) III - II currents.

These indirect effects are added to the currents due to the direct effects, and decrease the interference resistance of the circuits.

The resultant noise current in circuit II (the disturbance) will be:

at the near end: $I^{r}_{20} = I_{20} + I^{I}_{20} + I^{II}_{20} + I^{III}_{20}$,

at the far end: $I^{r}_{21} = I_{21} + I^{I}_{21} + I^{II}_{21} + I^{III}_{21}$,

where I_{20} and I_{21} are the direct effect currents;

I^{I}_{20} and I^{I}_{21} are the interaction currents due to load mismatches; Cu

I^{II}_{20} and I^{II}_{21} are the interaction currents due to internal nonuniformities of the cable;

I^{III}_{20} and I^{III}_{21} are the interaction currents through circuit III.

The values of the cross-talk attenuation at the near and far ends of the cable, B_0 and B_1, decrease correspondingly.

In addition to decreasing the cross-talk attenuation, the indirect interaction distorts its frequency dependence, and upsets the reciprocity of the interaction of circuits I and II. While under direct interaction, the cross-talk attenuation between circuits I and II, $B_{I/II}$, numerically equals the cross-talk attenuation between II and I, $B_{II/I}$, in the presence of indirect effects, this equality is destroyed, and $B_{I/II} \neq B_{II/I}$.

Figure 6-32 shows the cross-talk attenuation between circuits I and II with and without indirect interaction.

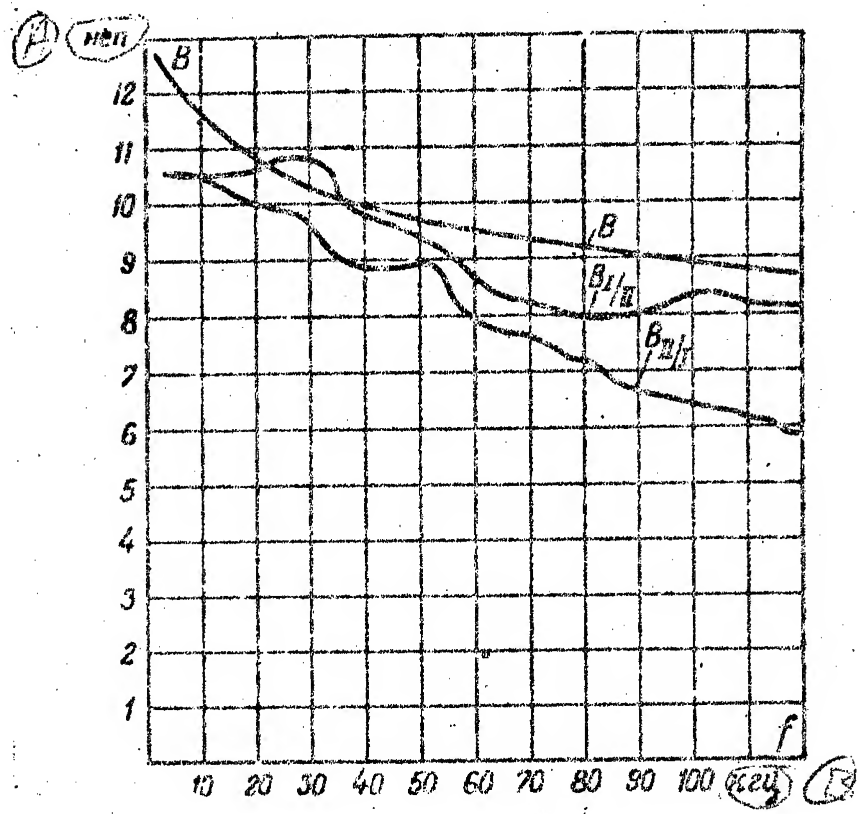


Fig. 6-32. Drop in cross-talk attenuation owing to indirect effects. B--direct cross-talk attenuation; $B_{I/II}$--and $B_{II/I}$--Cross-talk attenuation in the presence of indirect effects. A) nepers; B) kc.

6-12. INTERACTION IN COIL-LOADED CIRCUITS

Coil loading of cable circuits changes the ratio of the electromagnetic coupling coefficients K_0 and K_1 and has a substantial effect on the value of the cross-talk attenuation.

In Figs. 6-33 and 6-34, graphs are presented for the

variation with frequency of the electromagnetic coupling coefficients and the cross-talk attenuation at the near and far ends of styroflex cord insulated cables with and without coil loading.

From these graphs, it follows that:

1. The electromagnetic coupling coefficient at the near end, K_0, decreases as a result of coil loading, while the electromagnetic coupling coefficient at the far end, K_1, decreases. In the coil-loaded cable, the coefficients K_0 and K_1 are approximately equal in magnitude; this is explained by the fact that the wave impedance is 1.5 to 3 times larger than that of the nonloaded cable; because

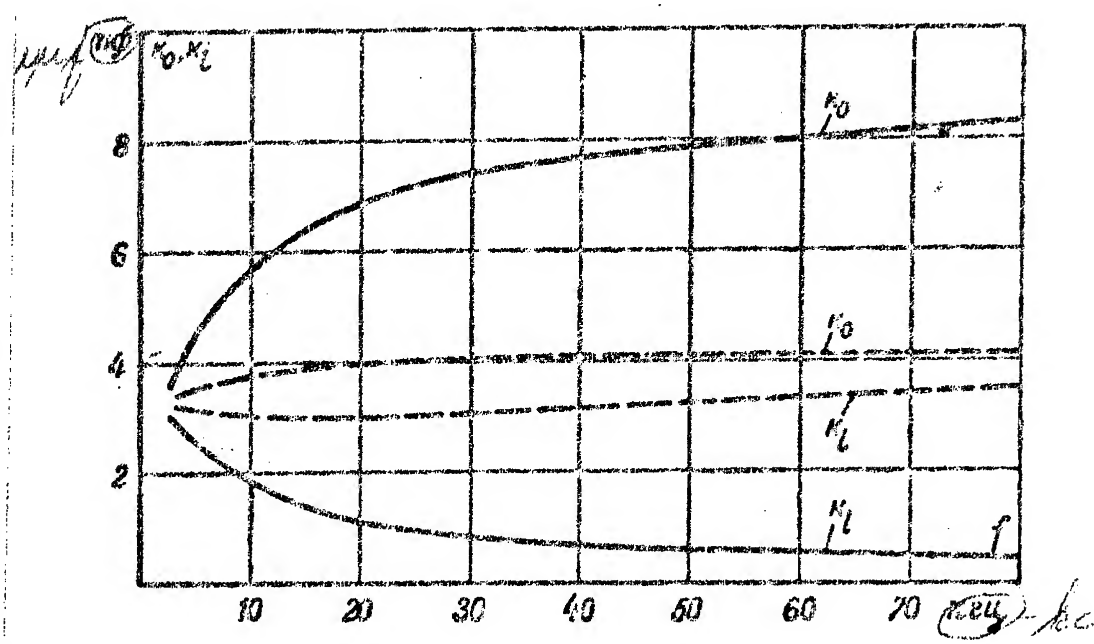


Fig. 6-33. Electromagnetic coupling coefficients for coil-loaded (- - - - - - -) and non-coil-loaded (————) cables.

of this, the effective inductive coupling $K_{\text{п}}$ is negligible, and all of the interaction is determined chiefly [sic] by the capacitive coupling.

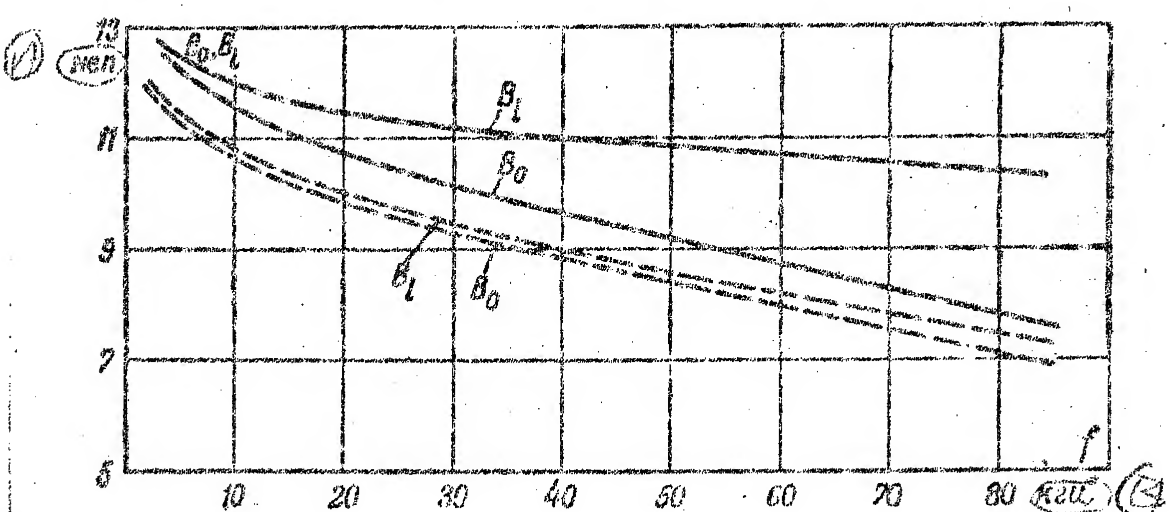


Fig. 6-34. The frequency of the cross-talk attenuation of coil-loaded (- - - - - - -) and non-coil-loaded (————) cables. A) nepers; B) kc.

As a result, the electromagnetic coupling coefficients at the near and far ends of the cable approach the value C_{12}, i.e.

$$K_0 \approx K_l \approx C_{12}.$$

2. The cross-talk attenuation of coil-loaded cables is less than that of non-loaded cables, and as the degree of loading is increased, the attenuation decreases, as is confirmed by the expression $B = \ln |2/\omega Z K|$.

Since the wave impedance Z_1 of a loaded cable is greater than the Z of the nonloaded cable, the difference

between the values B_1 and B of cross-talk attenuation in these cables is approximately

$$B - B_1 = \ln\left|\frac{Z_1}{Z}\right|^2; \qquad (6\text{-}50)$$

where quantities with the index 1 refer to the loaded cable.

The difference in cross-talk attenuation in the non-loaded and loaded cables is especially noticeable at the far end. The value of B_1 is 1.0 to 2.0 nepers less for the loaded cable. This is explained by the fact that in the nonloaded cable, since the capacitive and inductive couplings at the far end are about the same, the couplings tend to cancel, and the quantity K_1 is comparatively small.

In loaded cables, the capacitive and inductive couplings are no longer equal (the value of C_{12} is considerable more than M_{12}/Z^2), which raises the electromagnetic coupling coefficient K_1 and accordingly lowers the cross-talk attenuation. The higher the magnitude of the square of the ratio of the absolute values of the wave impedances of the loaded and nonloaded circuits, $(Z_1/Z)^2$, the more this is true.

Our conclusions are completely supported by the experimental data obtained as a result of measuring an experimental section of loaded (z_1 = 700 ohms) and non-loaded (z = 250 ohms) line, 30 km long (Table 6-18).

The difference in the cross-talk attenuation of loaded and nonloaded loops amounts on the average to 0.8 to 1.2 nepers, which sets severe requirements for the symmetry of cable lay-up and the electromagnetic coupling coefficients of loaded cables. It is especially important to obtain small capacitive coupling coefficients. Thus, although the existing standards for low-frequency cables allow C to equal 40 to 160 $\mu\mu$f per shipping-length section, for cables multiplexed over a band width of up to 60,000 cps, C should not exceed 5 to 10 $\mu\mu$f per shipping-length section. It is difficult to provide this kind of balance under industrial conditions, and therefore trunk cables are specially balanced while they are being assembled and laid.

Table 6-18. Cross-talk attenuation (in nepers), measured between circuits of a 30-km long cable. A) Frequency, kc; B) Cross-talk attenuation at the near end, B_{01}; C) without coil loading; D) with coil loading; E) Interference resistance at the far end, B_{12}; F) without coil loading; G) with coil loading; H) Attenuation of the circuit itself, $\beta \cdot n \cdot s$; I) without coil loading; J) with coil loading.

Table 6-18

Частота, *кгц*	1	2	3,5	5,5	7,5	9	10,5	13	20
Переходное затухание на ближний конец B_{on}									
а) без пупинизации	9,5	9,4	9,8	9,8	8,8	8,7	8,9	—	7,3
б) с пупинизацией	9,8	8,4	8,4	6,5	7,7	8,1	7,1	—	—
Защищенность на дальнем конце B_{l2}									
а) без пупинизации	10,7	10,6	10,0	9,0	—	—	—	—	—
б) с пупинизацией	9,8	9,3	8,7	8,8	—	—	—	—	—
Собственное затухание цепи $\beta \cdot n \cdot s$									
а) без пупинизации	1,9	2,6	3	3	—	—	—	—	—
б) с пупинизацией	1,0	1,4	2,0	2,5	3	—	—	—	—

6-13. FUNDAMENTAL CONCEPTS OF CABLE LAY-UP

The interaction of cable circuits and the electromagnetic coupling coefficients depend upon the configuration of the current-carrying conductors, which is determined by the method by which they are twisted in the cable, and upon irregularities in manufacture (varying conductor diameter, insulation which is not uniform, etc.); these are factors which can almost never be allowed for in advance.

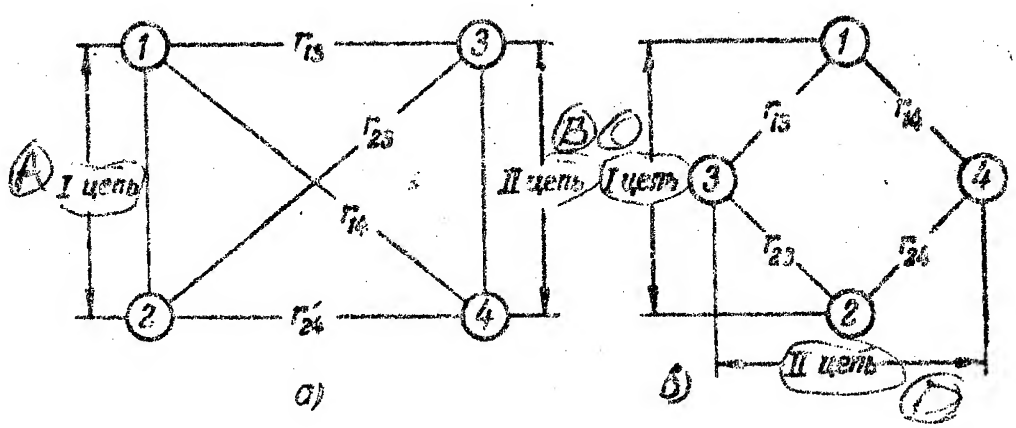


Fig. 6-35. Configuration of disturbing and disturbed circuits. a) General lay-up; b) star quadding. A) circuit I; B) circuit II; C) circuit I; D) circuit II.

When conductors are laid up into groups (pairs, starquads, double pairs, double starquads), and the groups are laid up into layers and into a common core for a cable there arises the problem of decreasing the electromagnetic couplings and interactions between the cable circuits.

The following types of coupling exist in cable circuits:

a) coupling within the group (coupling between the circuits of one or another group),

b) adjacent coupling (coupling between the circuits of different groups, which are located within one layer),

c) interlayer couplings (coupling between the circuits of groups that are located in different layers).

The capacitive and inductive coupling coefficients are expressed in terms of the distance between the disturbing (1-2) and disturbed (3-4) circuits by the following relationships (Fig. 6-35a)

inductive coupling

$$m = N_1 \ln \frac{r_{14} r_{23}}{r_{13} r_{24}}, \tag{6-51}$$

capacitive coupling

$$k = N_3 C_{12} C_{34} \ln \frac{r_{14} r_{23}}{r_{13} r_{24}}, \tag{6-52}$$

where r is the distance between conductors with the corresponding indices;

N_1 and N_2 are coefficients of proportionality, depending upon the insulating medium;

C_{12} and C_{34} are the capacitances between the conductors having the corresponding indices.

It is clear from Formulas (6-51) and (6-52) that the capacitive and inductive couplings are proportional to one another; therefore it is possible to decrease them simultaneously by using an appropriate configuration of the wires.

There will be no capacitive and inductive coupling when the following condition is fulfilled

$$\ln \frac{r_{14} r_{23}}{r_{13} r_{24}} = 0. \qquad (6\text{-}53)$$

In order for this condition to be fulfilled it is necessary that

$$\left.\begin{array}{l} r_{14} = r_{13} \text{ и } r_{23} = r_{24} \\ \text{or} \\ r_{14} = r_{24} \text{ и } r_{23} = r_{13} \end{array}\right\} \qquad (6\text{-}54)$$

It is also sufficient that the following equality holds

$$r_{14} r_{23} = r_{13} r_{24}. \qquad (6\text{-}55)$$

It is very easily seen that this condition is automatically satisfied by spiral twisting, where the disturbing circuits (conductors 1-2) and the disturbed circuit (conductors 3-4), are located on mutually perpendicular axes (Fig. 6-35b).

Since in this case the distances are equal ($r_{13} = r_{14}$) wire 1 has the same effect on both wire 3 and wire 4. Thus in wires 3 and 4 the currents due to a single source of interaction will flow in opposite directions, and the resultant noise currents in circuit 3-4 will be zero. Wire 2 also has no effect upon circuit 3-4, as this condition holds independently of the length of the lay.

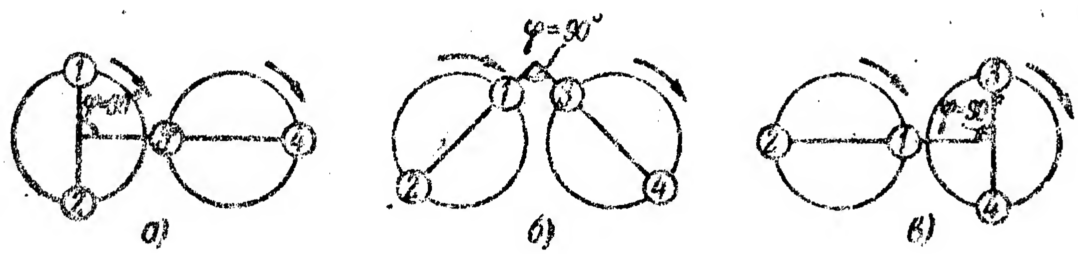


Fig. 6-36. Lateral symmetry method of layup. a) First configuration; b) second configuration; c) third configuration

In principle spiral twisting of the conductors of a quad guarantees the absence of intragroup coupling (between circuit 1-2 and circuit 3-4). When this type of layup is used coupling within the group can occur only where there

are variations and nonuniformities in the manufacturing characteristics, and cannot be due to the length of lay of the quad.

Coupling between spiral quads (adjacent coupling) depends to a large degree on the relationship of the quad lays.

In all the remaining configurations (double pair, double spiral quad, and paired groupings), the distance between the conductors of the disturbing and disturbed circuits varies continuously along the cable, and in order to achieve the minimum interaction special matching of the lays is required. This is true for both interaction with any group and between groups.

There are two possible methods for matching the lays and fulfilling the condition $r_{13}r_{24} = r_{14}r_{23}$: 1) the lateral symmetry method and 2) the longitudinal symmetry method.

Lateral symmetry of conductors is achieved by twisting the disturbing and disturbed circuits with the same lay but with a 90° phase shift.

As can be seen from Fig. 6-36, in this case at any given moment circuits I and II are located on mutually perpendicular axes, which guarantees that the equality $r_{13}r_{24}=$

$= r_{13}r_{23}$ will hold, and the interaction between the circuits will be reduced to a minimum. This method has not received wide application since it is exceptionally inconvenient in the manufacturing process and useless when a large number of circuits are involved. In practice it is only used for matching the lays of pairs in double pair groups. Longitudinal symmetry can be achieved by choosing different lays which are specially matched to each other.

The notion of matching the lays and the effectiveness of such a special selection can be clarified by the following example.

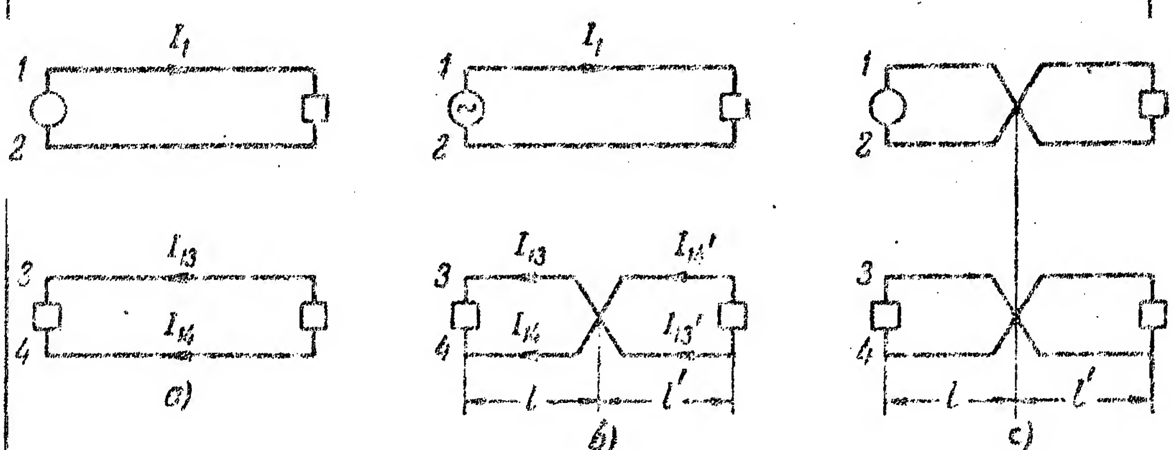


Fig. 6-37. Effectiveness of circuit transposition. a) Untransposed circuit; b) transposition has a positive effect; c) transposition gives no effect.

Fig. 6-37a shows two parallel circuit configurations: the disturbing (conductors 1-2) and disturbed (conductors 3-4). The current $\dot{I}_1$ in wire 1 induces the noise currents $\dot{I}_{13}$ and $\dot{I}_{14}$ in wires 3 and 4. Since wire 3 is located nearer to wire 1 than wire 4, the current $\dot{I}_{13}$ is larger than the current $\dot{I}_{14}$, and there will be a difference current $\dot{I}_{13}-\dot{I}_{14}$ in the second circuit.

The current $\dot{I}_2$ which flows in the second circuit acts in the same way causing a compensating current $\dot{I}_{23}-\dot{I}_{24}$. This compensating current appears in the receiving equipment of the circuit II in the form of interference.

Let us consider the interaction of the circuits, where one of the circuits, for example II, has been twisted at the center.

As Fig. 6-37b shows, in this case a current $\dot{I}_{13} + \dot{I}'_{13}$ will flow in wire 3, while in wire 4 there will flow a current $\dot{I}_{14} + \dot{I}'_{14}$; in this case in section $\underline{1}$ the current in wire 3 (I_{13}) will be greater than the current in wire 4 (I_{14}), while in section $\underline{1}'$, conversely, the current in wire 3 (I'_{13}) will be less than the current in wire 4 (I'_{14}).

Since the transposed sections are identical ($\underline{1} = \underline{1}'$), the total noise currents in wire 3 ($\dot{I}_{13} + \dot{I}'_{13}$)

and in wire 4 $(\dot{I}_{14} + \dot{I}'_{14})$ will be equal, but will be directed against one another, and therefore they will have no effect at the receiving equipment.

In this manner the noise currents in wires 3 and 4 which were caused by the current $\dot{I}_2$ flowing in wire 2, cancelled out.

Consequently interchanging the wires in one of the circuits (transposition) in general eliminates interference effects between the circuits.

This effect following upon transposition is based upon the equality of the currents flowing in sections 1 and 1'; however under actual conditions, owing to the attenuation of the currents along the line, the currents at the sending end are larger than those at the receiving end $(I_{13} > I'_{14}$ and $I_{14} > I'_{13})$ and the currents flowing in conductors 3 and 4 are never completely cancelled. A compensating noise current $(\dot{I}_{13} + \dot{I}'_{13}) - (\dot{I}_{14} + \dot{I}'_{14})$ will flow in the receiving equipment of circuit II.

If the circuit is transposed at several points rather than at one point the effect will be still better. The more frequently the conductors of the circuit are transposed the less difference there will be between the magnitudes of the currents of adjacent sections and the less the interference between the circuits.

It is easily seen that if both circuits are transposed at a single point rather than just one circuit the crossed portions neutralize each other and the effect of transposition disappears (Fig. 6-37c).

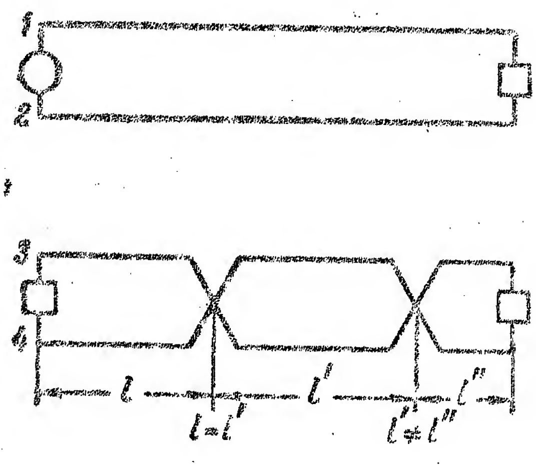


Fig. 6-38. Unequal length.

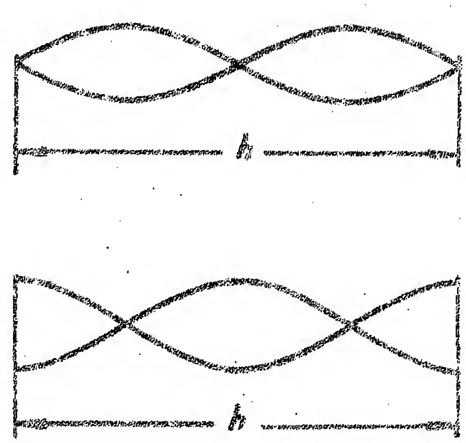


Fig. 6-39. Lay of the twist.

Thus crossing gives a positive effect only when the transposed circuits have different lays (the lay of the transposition is the distance between the adjacent points of transposition).

It should be noted that if the transposition lays of the circuit are not all equal a section of line not equal to the others in length will be formed (Fig. 6-38); this causes additional interaction between the circuits; the greater the length of the section the greater the interaction.

It is clear from Fig. 6-38 that if the noise currents cancel in sections l and l', then the unequal length l" acts as a nontransposed circuit and causes interaction between the circuits.

What has been said above leads to the conclusion that transposition is effective only for specially calculated lays differing for different circuits and reducing to a minimum inequalities in the length of sections of the circuits; in this case the more frequent the transpositions the greater the effect.

In principle the layup of the cable is similar to transposition, except that the latter is accomplished by interchanging (transposing) conductors at a point, while cable layup takes the form of a uniform distribution of the conductors along the length of the cable.

Transposition is chiefly used on aerial communication lines and for powerful currents; however its basic laws are correct for cable layup as well.

Each cable circuit is laid up with a different lay. By a lay h we mean the length in which the insulated conductor circuit or strand describes a complete circle about the axis about which twisting takes place (Fig. 6-39). The effect of the twist is greater the smaller the lay h.

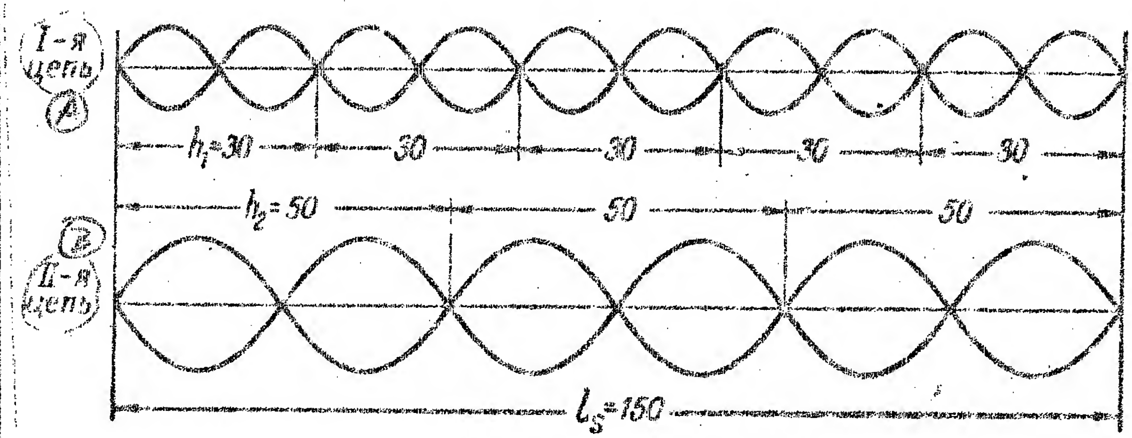


Fig. 6-40. Interference resistant section of two circuits. a) First circuit; b) second circuit.

The selection and matching of lays for different circuits and cable strands is carried out on the basis of sections which are called symmetric or protected sections. The protective section l_S is a section of cable along which a complete cycle of noise protection has been carried out for the cable circuit.

The protected section is connected with a lay by the following relationship

$$l_S = \frac{h_1 h_2}{B}, \qquad (6\text{-}56)$$

where B is the largest common denominator of h_1 and h_2.

For example if there are two circuits twisted with a lay $h_1 = 30$ mm and $h_2 = 50$ mm, the largest common denominator is $B = 10$ and

$$l_S = \frac{30 \cdot 50}{10} = 150 \text{ мм.}$$

As has been shown on Fig. 6-40 within the section $l_S = 150$ mm a complete cycle of interference protection (symmetricizing) has taken place between the two circuits considered.

Within the cable length l_S the configuration of the conductors in the pairs with respect to each other and the distance between the conductors r_{13}, r_{14}, r_{23}, r_{24} as well, constantly changes; at the end of the cable section l_S the same conductor configuration is obtained as at the beginning of the section. At the second and all succeeding sections of the cable of less length l_S the conductor configuration is repeated. Therefore it is enough to consider one single cycle of interference protection in a section.

From Fig. 6-41 where the nature of the variation in the distances between the conductors of two circuits has been shown in cable section l_s, it is clear that these distances are functions of the cable length l, and that the function $r_{14}(l)$, taken over the section l_1, shows the same variation as the function

$r_{24}(l)$ taken along the section l_2. In the interval l_1, the function $r_{13}(l)$ is equal to the function $r_{23}(l_2)$. in the interval l_2. At the center of the protected section l_s at the point A the distance between the conductors changes.

Consequently in a cable section l_s long the relationship

$$r_{14}(l)r_{23}(l) = r_{24}(l)r_{13}(l), \qquad (6\text{-}57)$$

between the functions holds when the logarithm of the fraction $r_{14}r_{23}/r_{24}r_{13} = 0$ and there will be no capacitive coupling k or magnetic coupling m, while the interference between circuits I and II will be reduced to a minimum; so for lateral symmetry the condition of the absence of capacitive and inductive coupling holds at any point of the cable cross section, while for longitudinal symmetry it occurs only at a cable length equal to the protection section l_s.

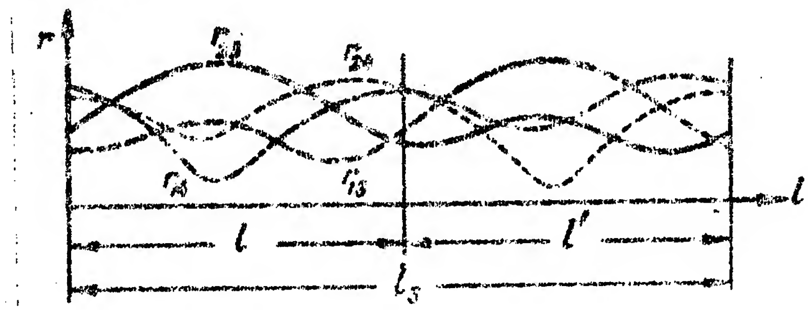


Fig. 6-41. Variation of the distance between the conductors of two circuits.

It is evident that in the ideal case where the circuits are to be protected from interference, the shipping length of the cable should contain a whole number of protected sections. In practice this requirement cannot be realized either in the manufacture, installation or assembly of the cable. Therefore it is desirable to keep the leftover unequal sections as short as possible. In order to do this the length of the protected section l_s is chosen as short as possible, so that when the cable is cut at any point the unequal section which results will be relatively short.

Decreasing the protection section also involves decreasing the lay h of the cable which is also advantageous from the electrical point of view since the less h is the greater the effect of twisting and the higher the noise resistance of the circuit. However there are manu-

facturing and technological considerations involved in using a very short lay since the shorter the lay h the greater the volume of the cable and the longer the actual length of the conductors.

At present for the sake of compromise long distance communications cables use a spacing on the order of 100 to 300 mm.

In high-frequency cables the lays of all cable groups must be matched; this is especially important for adjacent groups lying within one layer.

The matching of lays is also very important in the remaining cable groups however, since interference occurs in distant groups of the cable as a result of inductive coupling.

Capacitive coupling occurs only between the closest groups, since capacitive coupling between more widely separated groups is negligibly small. Capacitive coupling between two groups separated by a third have no importance even if their lays are not matched to each other. The intermediate group in the cable acts as an electrostatic screen to capacitive interaction, and absorbs the interfering electric field. Thus in low-frequency cables in which circuit interaction is caused in practice by capacitive couplings it is possible to match the lays of adjacent

groups of the cable alone. Thus it is sufficient to take two different mutually matching lays and alternate them. Thus for example in a layer having ten groups — five odd numbered groups are twisted with lay h_1, and the remaining even numbered groups, with a lay h_2.

For the odd numbered groups it is necessary to have a third matched lay in the layer h_3 (for the last group).

In addition to the direct interaction between circuits which is basically caused by direct capacitive coupling between the circuits and the zinc sheath there also occur interference effects to ground. The symmetry of the circuits with respect to ground is somewhat improved by the choice of matching lays, but unbalances still remain and it is necessary to attack them in the process of manufacturing the cable and in assembling the line.

When there are shielding groups in a cable it is necessary to keep in mind that the shield completely localizes capacitive interaction between the circuits alone. The interfering magnetic field is only partially decreased by the shielding shell, and so it is necessary to match the lays for these groups as well. It should be noted that where there are shielding groups in the cable the balance of the circuits with respect to ground is somewhat worse and the effect of matching the lays is somewhat increased.

In order to eliminate the harmful effect of a shield, the shielding circuit should be located either in the center of the cable, or in a peripheral layer.

In long distance communications cables, intended for high-frequency multiplexing, in which the role of inductive coupling is rather great, further intermatching of the groups is carried out.

The design and matching of lays is carried out with the aid of the following formulas:

a) for circuits with pair layup (P), located in a single layer

$$\frac{h_1}{h_2} = \frac{2v \pm 1}{2w}; \qquad (6\text{-}58)$$

b) for groups with spiral twisting (Z), located in one layer

$$\frac{h_1}{h_2} = \frac{4v \pm 1}{2w}, \qquad (6\text{-}59)$$

where h_1 and h_2 are the lays of matched groups;

v and w are any whole numbers greater than zero;

c) for pair and group twisting, DP (lying within a single layer)

$$
\left.
\begin{aligned}
h_1(a) &= h_1(\text{б}), \\
h_2(a) &= h_2(\text{б}), \\
\varphi_1(a) - \varphi_1(\text{б}) &= \frac{\pi}{2}, \\
\varphi_2(a) - \varphi_2(\text{б}) &= \frac{\pi}{2},
\end{aligned}
\right\} \qquad (6\text{-}60)
$$

$$
w_1 h_1 = \frac{2w_1 \pm 1}{2} h_2 = w_2 h_1(a) = \frac{4w_2 \pm 1}{4} h_2(a). \qquad (6\text{-}61)
$$

where h_1 and h_2 are the corresponding lays of the pair in the DP group;

$h_1(a)$; $h_1(b)$; $h_2(a)$; $h_2(b)$ are the corresponding lays of the conductors in the pair. The numeral designates the number of the DP group, the letter the number of the pair in the group. Thus, for example, $h_1(a)$ is the lay of the first pair of the first group;

$\varphi_1(a)$; $\varphi_1(b)$; $\varphi_2(a)$; $\varphi_2(b)$ are the corresponding angles between the axis of the pairs and the axis of abscissas. Thus, for example, $\varphi_1(a)$ is the angle between the axis of the first pair of the first group and the axis of abscissas.

It follows from expressions (6-58) to (6-61) that for twin pair twisting the pairs within the group are matched by the lateral symmetry method. They are wound with the same lay, and the angle between the axes of the pairs is 90 degrees.

Where it is necessary to match the lays of groups (pairs; spiral quads; double pairs), which are located in different layers of the cable, then in addition to the relationships set forth above, it is necessary to ensure that the following additional condition holds:

$$h_1 w = \frac{H_1 H_2}{H_1 \pm H_2} t, \qquad (6\text{-}62)$$

where H_1 and H_2 are, respectively, the lays of the first and second layers;

$\underline{w}$ and $\underline{t}$ are any integers greater than zero.

This places a definite restriction on the lays of the layers.

Formulas given above permit the calculation and matching of lays only for any two groups h_1 and h_2 or layers H_1 and H_2. In actuality a cable consists of a very large number of groups and all of them must be protected from interference. Ordinarily an arbitrary lay is assigned to the first group and different values of $\underline{w}$ and $\underline{v}$ are taken, a

considerable number of lays are calculated, and corresponding lays are chosen for each group.

By way of example let it be required to select the lays for a 4 by 4 spiral quad cable.

Since the lays of spiral groups must lie within limits of 100-300 mm, we take for the lay of the first pair 200 mm, and using the formula $h_1/h_2 = (4v \pm 1)/w$, we calculate the acceptable lays of the other three groups. The results of the calculation of the lays for values of v and w from 1 through 4 are given in Table 6-19.

TABLE 6-19.

Calculation of the Lays of Spiral Groups.

v \ w	1	2	3	4
1	160	320	480	640
2	89	178	266	355
3	61,5	121	184	245
4	47	94	140	186

Any of the lays given will provide proper protection against interference for the cable circuits, but in order to decrease the effect due to unequal lengths it is necessary to choose that lay h for which the protected section l_s is shortest. In addition it is desirable that all of the

lays chosen fall within specific limits, for example 100 to 300 mm.

In 4 by 4 styroflex-insulated cable the following lays are used: $h_1 = 200$ mm; $h_2 = 160$ mm; $h_3 = 175$ mm; $h_4 = 125$ mm. The tolerance is ± 5 mm.

6-14. PRINCIPLES FOR ORGANIZING LONG DISTANCE COMMUNICATIONS OVER TRUNK CABLES

Let us consider the contemporary principles by which long distance communications are set up over wire circuits and, comparing them, set up fundamental concepts for designing cables.

As is known cables may be utilized in long distance communications in the following ways: 1) for low and high frequency transmissions; 2) in two-wire and four-wire systems; 3) in single-cable and double-cable communications systems.

Low frequency communication takes in the 0-3000 cps band, and includes subtonal telegraphy (0-100 cps) and voice frequency telegraphy (300-3000 cps).

The high frequency region used for multiplexed cable communications begins at 5-6 kc and reaches 100 kc (for symmetric cables) and 1000 kc (for coaxial cables). This spectrum is used for telephone and telegraph

transmission in high-frequency channels; this was made possible by the creation of electron tubes and electric filters which enter into the receiving and transmitting equipment.

The development of methods for multiplexing the high-frequency band was occasioned by the desire for effective use of expensive cable lines and for obtaining the largest possible number of telephone and telegraph channels, and in order to satisfy the growing requirements for ever longer transmission distances.

The maximum distance for low-frequency voice-frequency communication over cable lines does not exceed 1000 km, since any further increase is limited by the two-way repeaters, the total number of which may not exceed 8 to 10 in the trunk.

This is explained by the fact that normal vacuum-tube amplifiers are used to amplify in one direction only, while the two-way repeaters must utilize two opposed amplifiers.

If they are connected as shown in Fig. 6-42 oscillation will occur (self-excitation), and the circuit will not operate.

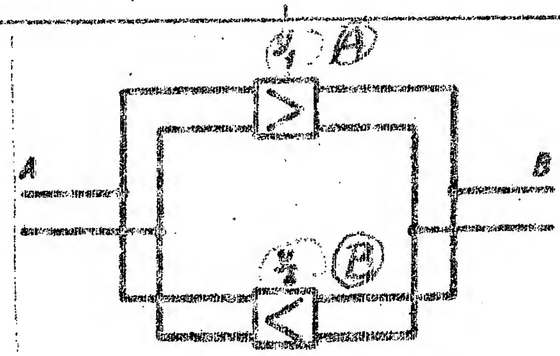


Fig. 6-42. Basic circuit of two-way repeater: A) amplifier 1; B) amplifier 2.

By using differential transformers such as T_1 and T_2 with the two-way repeaters, together with balancing circuits such as BK_1 and BK_2, it is in principle possible to eliminate the danger of oscillation and to provide stability and independent amplification of the currents transmitted in the appropriate direction (Fig. 6-43).

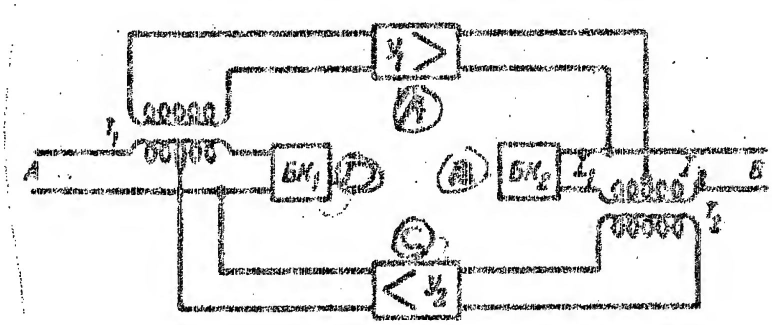


Fig. 6-43. Two-wire low-frequency communications system: A) amplifier 1; B) balancing circuit 2; C) amplifier 2; D) balancing circuit 1.

But such [illegible] (ideal) operation of a two-way repeater is possible only where the resistances of the balancing circuit equal the input impedance of the line. However when a balancing circuit is used it is not possible to reproduce exactly the input impedance of a line since its characteristics vary under the influence of different atmospheric factors, and consequently, there will be an unbalance in the system and as a result self-excitation of the amplifier is inevitable. The more amplifiers that are connected into the circuit, the more strongly they will interact and the greater the danger that oscillation will occur. In practice this limits the number of amplifiers connected in series in the circuit to 8-10 and accordingly limits the distance at which low-frequency communications can be carried on over a two-wire system. On the basis of stability consideration amplifiers used on low-frequency two-wire systems should not have more than 1.5 to 2 nepers gain.

An actual method for increasing the distance is the four-wire communications system. As may be seen from Fig. 6-43 in the two-wire system transmission in both the forward and reverse direction takes place over one pair of wires with the four-wire system four conductors are used for transmitting; one pair of conductors is used for com-

munications in one direction, another pair for communications in the other direction. Communications over an interurban cable link are commonly set up on the basis of a four-wire system. The system has not been widely applied to aerial links because of the necessity of using twice the number of conductors, and the difficulty of building multiconductor pole lines.

In cable links where the number of current-carrying conductors is relatively large, extra conductors are even provided which have all the characteristics required for use in a four-wire communications system.

As can be seen from Fig. 6-44 in a four-wire system the amplifiers to the forward and reverse directions are not interconnected, which excludes the possibility of oscillation. Thus in this system the distance of communication is not limited by the stability of the amplifiers. The high cost of the four-wire communications system is a serious drawback at voice frequencies; for this system the cable must have double the capacity of the two-wire system.

The four-wire system yields good technical and economic results only when high-frequency multiplexing is used with the cable links. This is explained by the following structural features of high-frequency multichannel links.

Regardless of the system of communications a low-

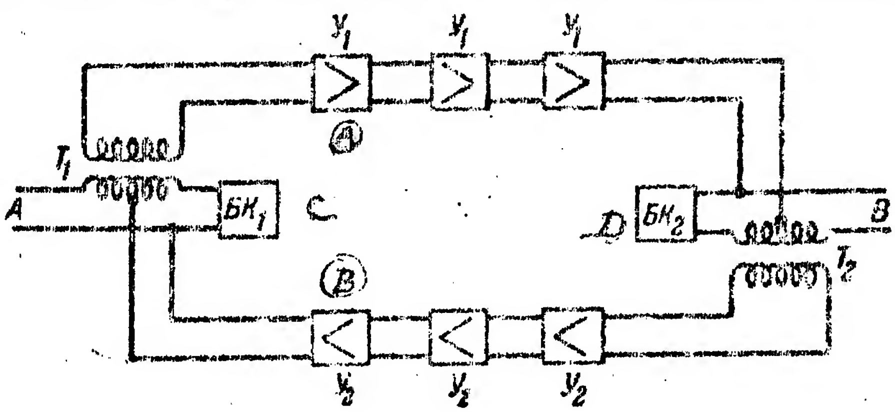


Fig. 6-44. Four-wire low-frequency communications system. A) Amplifier 1; B) amplifier 2; C) balancing circuit 1; D) balancing circuit 2.

frequency telephone transmission occupies exactly the same voice-frequency band in the forward and reverse directions (300-3000 cps), with limited distance, as was shown above.

In a high-frequency link the attenuation in the repeatered sections is considerably above that found in low-frequency links; it reaches 6-7 nepers. It is clear that under these conditions two-way repeaters with balancing circuits cannot operate with satisfactory stability since it is nearly impossible to match the circuits as accurately as would be required for a gain of this order. Thus for high-frequency links transmission in opposite directions is accomplished by using different portions of the frequency spectrum or by using independent pairs of conductors.

In setting up long-distance high-frequency communication over a two-wire system the frequency spectrum is divided into two parts: a lower part and an upper part. The lower part of the spectrum is utilized for transmitting in one direction and the upper part for transmitting in the other.

For dividing the transmissions in the reverse and forward directions, and to preclude the possibility of oscillation at the input and output of each amplifier, directional (separating) filters are used (Fig. 6-45).

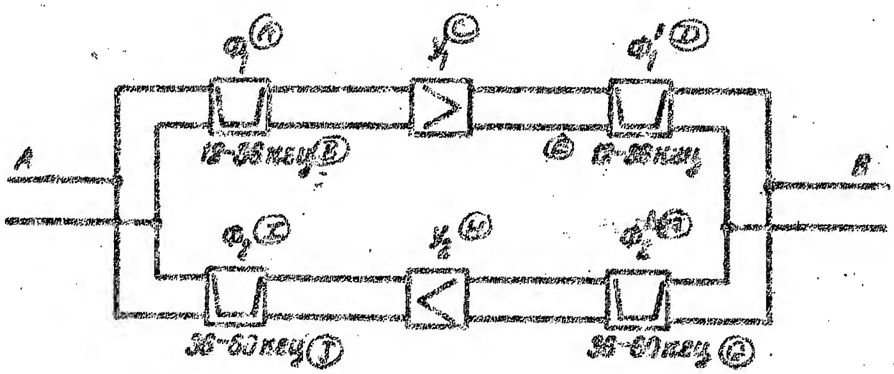


Fig. 6-45. High-frequency communications over a two-wire system using separation filters. A) Filter$_1$; B) 12 to 36 kc; C) amplifier 1; D) filter'$_1$; E) 12-36 kc; F) filter'$_2$; G) 36-60 kc; H) amplifier 2; I) filter 2; J) 36-60 kc.

In the four-wire system the forward and reverse directions use exactly the same frequency band but as was shown above different pairs of conductors are used for the

forward and reverse transmissions and separation filters are not required.

This latter fact is the chief advantage of the four-wire high-frequency communications system since it considerably simplifies the amplifying equipment (Fig. 6-46).

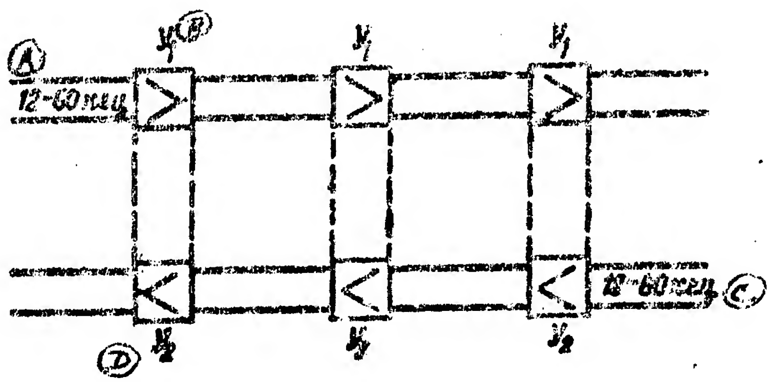


Fig. 6-46. High-frequency communication over a four-wire system. A) 12-60 kc; B) amplifier 1; C) 12-60 kc; D) amplifier 2.

It should be noted that as far as the number of channels are concerned the two-wire and four-wire systems of high-frequency communication are the same. This idea is graphically illustrated by the following example.

Let us consider a twelve-channel multiplexed system in a symmetrical cable; the system uses a range of from 12 to 60 kc with a nominal 4 kc per channel.

In the two-wire system the first half of the

spectrum (12-36 kc) is reserved for communication in one direction, and the second half (36-60 kc) for transmission in the other direction. As a result in one pair of wires six two-way transmissions take place. In order to obtain 12 transmissions it is necessary to use two pair of wires. In the four-wire system one pair carries 12 communications (12-60 kc) in one direction while the second pair using the same frequency range (12-60 kc) carries 12 communications in the reverse direction. The total number of communications taking place over two pairs of wires is exactly 12 for both the four-wire system and the two-wire system (Fig. 6-47).

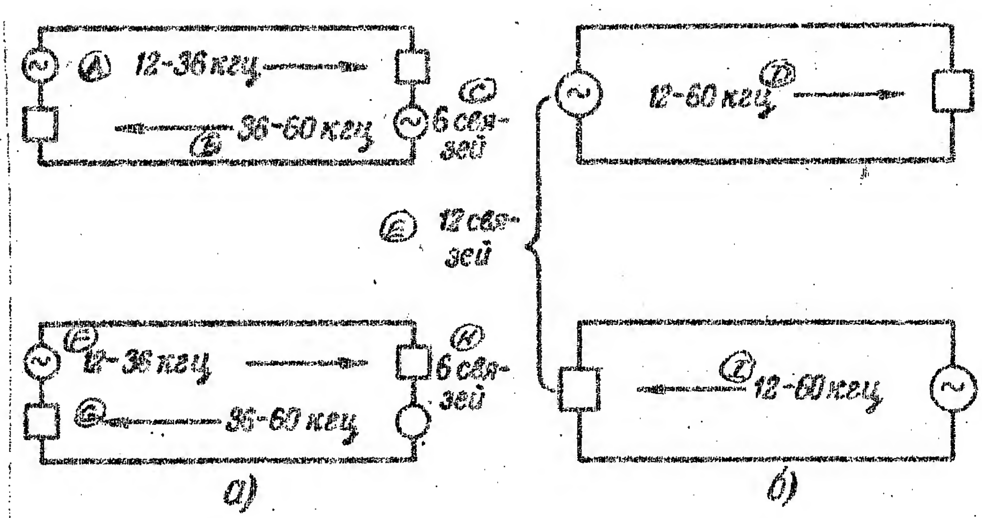


Fig. 6-47. Two- and four-wire high-frequency communications systems compared as to the number of communications channels. a) Two-wire communications system; b) four-wire communications system. A) 12-36 kc; B) 36-60 kc; C) six communications; D) 12-60 kc; E) 12 communications; F) 12-36kc; G) 36-60 kc; H) six communications; I) 12-60 kc.

On the basis of what has been presented we may say:

1. Low-frequency communication is limited in range with the cable links used in a two-wire system and is not suitable for long line intercity communication.

2. Utilization of cables in four-wire systems provides the necessary range of communication, but is not economically feasible for low-frequency communication.

3. The best way to organize long distance intercity communication over a cable line is carrier frequency multiplexing using a four-wire system.

In order to compare the two- and four-wire communications systems thoroughly, it is also necessary to consider the interaction of circuits under each of the systems.

Figure 6-48 gives a diagram of the interaction of circuits in the two-wire and four-wire systems. In the two-wire system, a frequency band (for example, 12-36 kc) is transmitted in the direction A-B, while in the reverse direction from B to A, another band (for example, 36-60 kc) is transmitted.

Since in any given direction exactly the same frequency spectrum is transmitted over all the circuits of the cable (for example, from A to B, frequencies from 12 to 36 kc), the most dangerous interferences occur at the

far end. There is no noise energy at the near end, since the filtering equipment does not pass this frequency band.

In the four-wire system, in both the direct (A—B) and reverse (B—A) directions exactly the same frequency band is transmitted; 12-60 kc in the example given above. Therefore in this case interaction can occur between the circuits at both the near and far ends of the cable.

As has been shown above, the cable crosstalk attenuation at the far end B_1, is greater as a rule, than that at the near end B_0, and for cables operated under carrier multiplexing, it is extremely difficult to attain the required value of interference resistance at the near end of the circuits.

Consequently, from the point of view of the interaction between the cable circuits, the two-wire carrier-frequency communications system operates under more satisfactory conditions in comparison to the four-wire system.

In order to increase the interference resistance of the circuit and exclude undesirable interaction at the near ends in a four-wire system a two cable communications setup can be used. In this case the direct and reverse circuits are located in separate cables (the A—B direction circuits in cable No. 1, and the B—A direction

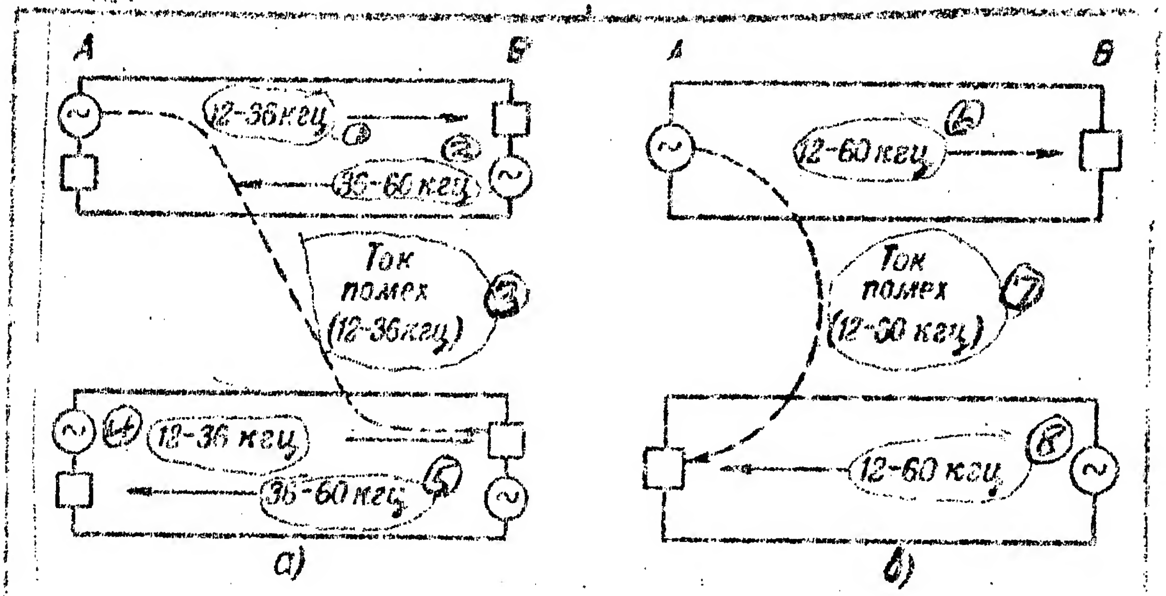


Fig. 6-48. Interaction in the various systems for using cables in long distance communication. a) Two-wire communications system; b) four-wire communications system. 1) 12-36 kc; 2) 36-60 kc; 3) noise current (12-36 kc); 4) 12-36 kc; 5) 36-60 kc; 6) 12-60 kc; 7) noise current (12-60 kc); 8) 12-60 kc.

circuits in cable No. 2).

In a two cable setup the interference resistance is determined by the interaction of the circuit at the far end of the line (as a rule, both cables are laid in the same trench).

It is also possible to separate the forward and reverse circuits by using electromagnetic shields.

As can be seen from Fig. 8-2 the A–B and B–A direction circuits are screened from each other by a dividing shield which is radial, grouped or circular.

With screened cables it is possible to achieve

high quality multiplexing using a single cable setup. Thus, for a four-wire system of carrier-frequency communications, it is necessary to use either cables with separating shields, or a two-cable setup.

At present when long distance communications systems are set up, two cable links are being used.

6-15. BALANCING OF CABLE LINKS

The balancing of cable circuits is the fundamental method for protecting them against external and internal interference, and providing high quality communications over long distances. It amounts to compensating the electromagnetic couplings acting in the cable in order to raise the interference resistance of the cables and the crosstalk attenuation as well.

Balancing is carried out both under factory conditions, and during the installation of cable lines.

In order to produce completely satisfactory cables, meeting the standards in the most important characteristic — the electromagnetic coupling coefficient — it is necessary to carry out an entire series of steps at the factory both in the process of preparing for production and in the actual manufacture of the cables.

The problem is to ensure that the following characteristics are maximized: the geometric symmetry of cable

pairs and quads, the uniformity of the basic materials (copper, paper, etc.) and electrical uniformity of the cable parameters (primarily R and C).

One of the basic conditions for decreasing the electromagnetic-coupling coefficients is the calculation and choice of matched lays for the cable circuits (see section 6-13).

If measurements performed on the finished cable show that any quad does not meet the standards with respect to the coupling coefficients then it must be balanced. The balancing standards are shown in Tables 6-11, 6-12 and 6-13.

Under factory conditions cables are balanced by the transposition method. The advantages of this method in comparison to the method of balancing by means of added capacitors are: a) only the capacitive couplings are balanced by using the capacitances, while transposition compensates for both capacitive and inductive coupling; b) the transposition method is more economical since it does not require balancing capacitors and does not cause local thickening of the cable.

As a rule, balancing is carried out in the middle of a shipping length section of cable. In order to do this the shipping length cable section is cut in two, the cor-

responding pairs and quads of the cable are balanced, and then the conductors are spliced and the lead sheathing restored. Cables which are shipped in large lengths (for example submarine cables) should be balanced by a concentrated balancing method.

When being laid low-frequency cable trunks are balanced during the installation of the connections. Cables are balanced in sections called balancing lengths. The balancing length of loaded cables equals the loading length and in the majority of cases equals 1.7 km. In non-loaded cables the balancing length can range up to 4 km.

Since in low-frequency cables capacitive coupling predominates, the required balancing effect can be obtained by transposition and by connecting in additional capacitors.

As a rule balancing is achieved by a combination of methods. First balancing is carried out by transposition, and then the remaining unbalance is remoted by connecting additional capacitors.

Transposition balancing is carried out at each connection, and special capacitor connectors are used. In order to achieve the best results, balancing is carried out according to a specific scheme, and in a definite order.

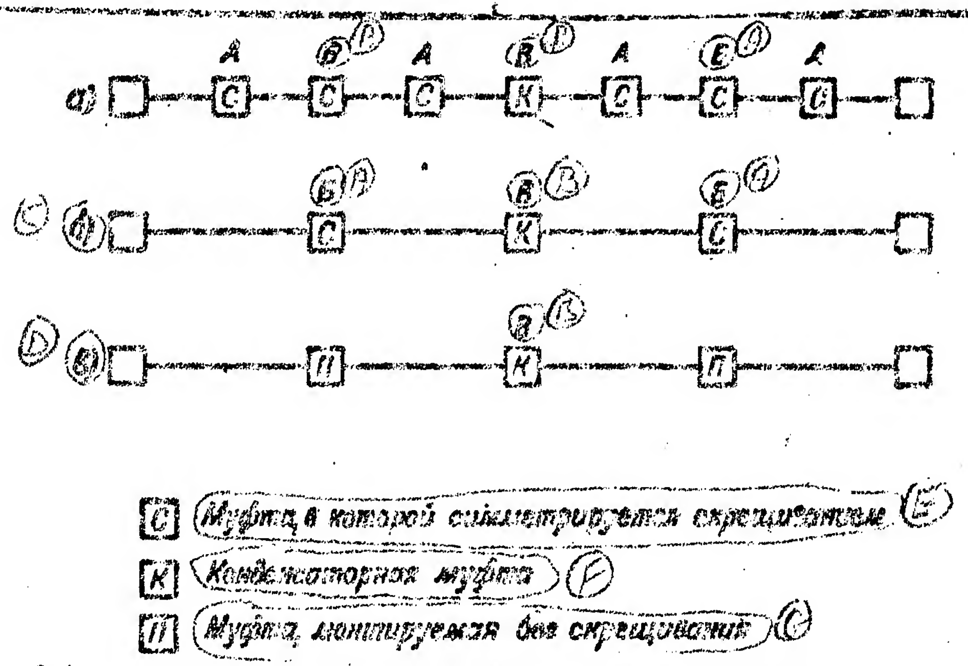


Fig. 6-49. Schemes for balancing a cable within a balancing length. a) Seven point scheme; b) three point scheme; c) single point scheme. A) Б; B) С; C) b; D) c; E) transposition-balancing connection; F) capacitor connection; G) connection made without transposition.

A scheme for balancing a cable within a balancing length (the so-called seven point scheme) is given in Fig. 6-49.

The cable is first balanced by transposition at connections A. The second stage of transposition is carried out at connections B, and the third and final stage at the capacitor connection C, where in addition to the transposition, balancing capacitors are inserted; with the aid of the latter the unbalance and coupling are reduced to acceptable values.

The capacitive coupling and unbalance coefficients in a balancing length of installed cable, at a frequency of 800 cps, should not exceed the values given in Table 6-14.

In many cases low-frequency cables may be balanced by using three point and even single point schemes (Fig. 6-49b and 6-49c). The fewer the balancing points, the simpler the assembly of the cable.

Regardless of the scheme used, the cable is balanced at connection C at first by transposition, and only later, where necessary, is the residual unbalance eliminated with the aid of balancing capacitors.

It should be kept in mind that what has been said above relates to balancing within the limits of a balancing section. Over the length of the cable it is necessary to work not with the coupling coefficients, but rather with the values of crosstalk attenuation. Therefore the interconnection of the individual balancing lengths is carried out on the basis of measurements of the crosstalk attenuation in the cable.

The peculiarities of balancing high-frequency cables are determined by the nature of interaction in high-frequency circuits. Here inductive couplings act in addition to capacitive couplings, along with the in phase

components; on the whole the coupling is complex (vector) in nature, and varies with frequency both in absolute value and in phase. High-frequency cable trunks are balanced in two stages.

The first stage amounts to common section-wise balancing, similar to that employed for low-frequency cables. The coefficients k and e are compensated by means of transposition and reduced by capacitors into the standard values given in Table 6-14.

The second stage is carried out after the repeatered section of cable has been assembled, and consists of concentrated balancing of the cable circuits; first a concentrated transposition is made, and then (where necessary) couplings are compensated by connecting lump equalizing elements to oppose the coupling. Both transposition as well as the connection of anti-coupling elements is carried out at one or two points of a repeatered section of the cable line.

A completely balanced repeatered section of line should satisfy the standards for interference resistance and crosstalk attenuation. For high-frequency circuits, the interloop noise resistance B_{12} (per repeatered section) should be, for all frequency bands used in practice, not less than 7.5 nepers.

6-15. BALANCING BY THE TRANSPOSITION METHOD

This method consists of compensating the unbalance of a quad in one shipping length of cable A, by the unbalance of a quad in another length B, by direct connection of the corresponding quads of two adjacent lengths of cable or by transposition of the conductors in the quads.

Coupling compensation is based upon the fact that for direct (straight) connection of the conductors, the coefficients of coupling for different lengths are added algebraically ($k^A + k^B$), while for transposition they subtract ($k^A - k^B$). Owing to the transposition the sign of the coupling coefficient is reversed.

If connected cable lengths A and B have coefficients of coupling which differ in sign, they should be transposed "directly." The total coefficients of the cable A + B will in this case decrease.

Thus for example: $k^A = -400$ μμf,

$k^B = 500$ μμf.

The total coefficient $k^{A+B} = k^A + k^B = 100$ μμf.

If the coefficients for cables A and B have the same sign, then it is desirable to connect them not "directly," but rather by transposing the conductors of one cable relative to the conductors of the other cable; here the total coefficients of the cable A + B will also de-

crease. Here the transposition will be most effective where the quads of cable A and of cable B are transposed with equal absolute values of the couplings coefficients. In this case the coupling coefficients will be equal in value but opposite in sign and will completely cancel out.

Thus for example: $k^A = 400$ µµf,

$k^B = 400$ µµf.

The total coefficient is $k^{A+B} = k^A - k^B = 0$.

There are eight possible transposition combinations when a single quad of cables A and B are joined.

Table 6-20 gives the operations involved in balancing by the transposition method, and the arrangements corresponding to them. The operations are represented by a three member symbol: the first member refers to the first circuit, the second to the second circuit, and the third member characterizes the phantom circuit.

A dot (·) designates a "direct" connection, while a multiplication sign (X) designates a transposed connection. Thus for example, the operation (·XX) indicates that the first circuit is "directly" connected, while the second and phantom circuits are transposed.

The various combinations of transposition give rise to various values of the resultant coefficients (k_1, k_2, k_3 and e_1, e_2, e_3).

For some sections A and B the values of the coefficients will add while for others they will subtract.

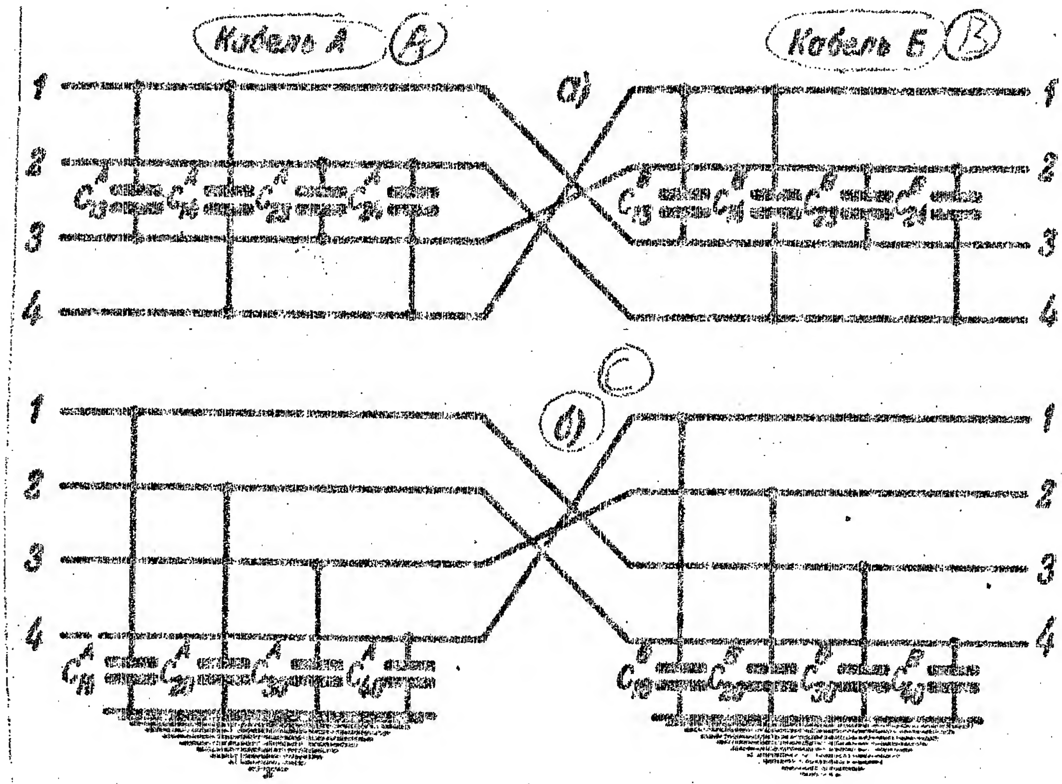


Fig. 6-50. Determination of the total coefficients for the cable A + B with the operation ·XX. A) Cable A; B) cable B; C) b.

Let us consider the resultant values of the coefficients k_1, k_2, k_3 and e_1, e_2, e_3, for example, for the seventh combination — the operation (·XX).

As shown in Fig. 6-50, the total direct capacitances of the quad of cable A + B, transposed according

to the given operation, will be:

$$c_{13}^{A+B}=c_{13}^{A}+c_{23}^{B}, \quad c_{24}^{A+B}=c_{24}^{A}+c_{14}^{B},$$

$$c_{14}^{A+B}=c_{14}^{A}+c_{13}^{B}, \quad c_{23}^{A+B}=c_{23}^{A}+c_{24}^{B},$$

$$c_{10}^{A+B}=c_{10}^{A}+c_{30}^{B}, \quad c_{20}^{A+B}=c_{20}^{A}+c_{40}^{B},$$

$$c_{30}^{A+B}=c_{30}^{A}+c_{20}^{B}, \quad c_{40}^{A+B}=c_{40}^{A}+c_{10}^{B}.$$

Substituting the value found for the total direct capacitances into the formulas for capacitive and balanced coefficients (k and e), we obtain values for the latter, expressed in terms of the coefficients for balanced cables A and B.

Thus for example for the coupling coefficient k_1

$$k_1^A=(c_{13}^A+c_{24}^A)-(c_{14}^A+c_{23}^A),$$

$$k_1^B=(c_{13}^B+c_{24}^B)-(c_{14}^B+c_{23}^B),$$

whence for the total coefficient

$$k_1^{A+B}=(c_{13}^{A+B}+c_{24}^{A+B})-(c_{14}^{A+B}+c_{23}^{A+B}).$$

Substituting in the values of the total direct capacitances, we obtain

$$k_1^{A+B}=(c_{13}^A+c_{23}^B+c_{24}^A+c_{14}^B)-(c_{14}^A+c_{13}^B+c_{23}^A+c_{24}^B)=$$

$$=[(c_{13}^A+c_{24}^A)-(c_{14}^A+c_{23}^A)]-[(c_{13}^B+c_{24}^B)-(c_{14}^B+c_{23}^B)]=k_1^A-k_1^B.$$

Table 6-20. Operations for Balancing by the Transposition Method. A) No.; B) operation; C) symbol; D) transposition scheme; E) cable A; F) cable B; G) value of resultant coefficient; H) notes; I) I — first circuit; J) II — second circuit; K) Ф — phantom circuit; L) • — direct connection; M) X — transposed circuit; N) Ф.

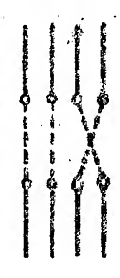
(A) № по пор.	(B) Оператор: (C) Обозначение I	II	Ф	(D) Схема скрещивания (E) Кабель А (F) Кабель Б	(G) Значение результиру... k_1	k_2	k_3
1	•	•	•		$k_1^A+k_1^B$	$k_2^A+k_2^B$	$k_3^A+k_3^B$
2	X	•	•		$k_1^A-k_1^B$	$k_2^A-k_2^B$	$k_3^A+k_3^B$
3	•	X	•		$k_1^A-k_1^B$	$k_2^A+k_2^B$	$k_3^A-k_3^B$
4	X	X	•		$k_1^A+k_1^B$	$k_2^A-k_2^B$	$k_3^A-k_3^B$
5	•	•	X		$k_1^A+k_1^B$	$k_2^A+k_3^B$	$k_3^A+k_2^B$
6	X	•	X		$k_1^A-k_1^B$	$k_2^A-k_3^B$	$k_3^A+k_2^B$
7	•	X	X		$k_1^A-k_1^B$	$k_2^A+k_3^B$	$k_3^A-k_2^B$
8	X	X	X		$k_1^A+k_1^B$	$k_2^A-k_3^B$	$k_3^A-k_2^B$

Table 6-20
Continued

...ющего коэффициента			Примечание (H)
e_1	e_2	e_3	
$e_1^A + e_1^B$	$e_2^A + e_2^B$	$e_3^A + e_3^B$	I — первая цепь (I) II — вторая цепь (J) Ф — фантомная цепь (K) · — соединение на прямую (L) X — скрещивание цепи (M)
$e_1^A - e_1^B$	$e_2^A + e_2^B$	$e_3^A + e_3^B$	
$e_1^A + e_1^B$	$e_2^A - e_2^B$	$e_3^A + e_3^B$	
$e_1^A - e_1^B$	$e_2^A - e_2^B$	$e_3^A + e_3^B$	
$e_1^A + e_2^B$	$e_2^A + e_1^B$	$e_3^A - e_3^B$	
$e_1^A - e_2^B$	$e_2^A + e_1^B$	$e_3^A - e_3^B$	
$e_1^A + e_2^B$	$e_2^A - e_1^B$	$e_3^A - e_3^B$	
$e_1^A - e_2^B$	$e_2^A - e_1^B$	$e_3^A - e_3^B$	

Consequently for the seventh operation of transposition the sign of the coefficient k_1^B is reversed and the coefficient k_1 for balanced cables subtract.

In like manner it may be shown that when the resultant coefficient e_1 is determined, the coefficients do not differ in sign for balanced cables, but add

$$e_1^{A+B} = c_{10}^{A+B} - c_{20}^{A+B} = (c_{10}^A + c_{30}^B) - (c_{20}^A + c_{40}^B) =$$
$$= (c_{10}^A - c_{20}^A) + (c_{30}^B - c_{40}^B).$$

The expression obtained may be given in the form:

$$e_1^{A+B} = e_1^A + e_2^B.$$

The remaining resultant coefficients k_2, k_3, e_2, and e_3 may be obtained by the same method.

From Table 6-20, where the results of calculation of all six coupling and balance parameters are given for a cable quad with differing combinations of transposition, it is clear that only with the first operation the resulting coefficients equal the sum of the coefficients of the balanced cables A and B. For all the remaining operations part of the resultant coefficients are expressed by the sum of coefficients A and B, and part by their difference.

In accordance with this it may be pointed out that the use of any operation for single coefficients is ef-

fective, while for others it is completely unsuitable. This is explained by the complexity of choosing an operation and transposition scheme, since it is necessary to choose a scheme which provides a decrease in the maximum number of resultant coupling and unbalance coefficients.

As an example let us choose an operation for balancing a cable by the transposition method for a cable not utilizing phantom circuits (which considerably simplifies the problem, since it makes it possible to operate solely with the coefficients k_1, e_1, e_2 and the first four transposition schemes).

Taking the measurement results given in columns 2 and 3 of Table 6-21 for the coupling coefficients and unbalance coefficients of balanced cables A and B, the choice of an operation is carried out as follows.

The resultant couplings (k_1, e_1, e_2) for all operations are computed and set down without signs in columns 4-7 of Table 6-21.

Наименование коэффициентов	Результаты измерения коэффициентов связи и асимметрии ($n\phi$)		Результирующие связи при различных операторах			
	Кабель А	Кабель Б	...	X..	.X.	XX.
1	2	3	4	5	6	7
k_1	—30	+40	10	70	70	10
e_1	+20	+20	40	0	40	0
e_2	+30	+20	50	50	10	10
Максимальное значение остаточной связи			50	70	70	10
Выбираемый оператор				XX.		

Table 6-21. Balancing by the Transposition Method (Without Taking Account of Phantom Links). A) Coefficients; B) results of measuring the coupling and unbalance coefficients (μμf); C) cable A; D) cable B; E) resultant coupling for the different operations; F) maximum value of the residual coupling; G) operation chosen.

The resultant coupling is determined with the aid of Table 6-21, in which it is shown whether the coefficients of cables A and B must be added or subtracted.

Thus for example for the operation (X..):

$$k_1^{A+B} = k^A - k^B,$$

consequently $k_1^{A+B} = -30 - (+40) = -70.$

At the bottom of Table 6-21 the maximum values of the couplings are given on the basis of each operation and the results obtained are analyzed. It follows from the table that if the conductors of the cable are connected "directly" (...), then the resulting couplings will amount to $k_1 = 10$; $e_1 = 40$; $e_2 = 50$ (column 4), while when transposition is carried out on the basis of the operation (XX.) the couplings are considerably decreased and equal respectively 10, 0 and 10 (column 7).

When the operations (.X.) and (X..) are used, poorer data are obtained than when the cables are joined "directly."

As a result for balancing of the cable quad under consideration we choose the operation (XX.).

In order to speed up the choice of a transposition operation in actual balancing, previously prepared tables are used. When these tables are available the choice of

an operation is simplified and reduces to noting down the sums of the coefficients of cables A and B in the free columns.

Table 6-22 illustrates the process of balancing a cable quad.

The results of measurements of the coupling and unbalance coefficients for cables A and B are given in columns 1, 2 and 8.

The absolute values of the coefficients are added and noted in the open uncrosshatched boxes. If the coefficients of cables A and B differ in sign, then the "minus" sign is used while if they have the same signs the "plus" sign is used.

The unbalance coefficients (e_1,e_2,e_3) are written in columns 4-7 and 11-14, arbitrarily reduced by 10 times. The reason for this is that there are more stringent requirements placed upon the coupling coefficients than with respect to the unbalance coefficients (see section 6-8). Then the maximum value of the sum is given in the bottom line of the table for each operation. The operation is chosen on the basis of the minimum sum to be found in the bottom line.

As may be seen from the table in the example given the minimum sum corresponds to the operation (.X.), which

is used for balancing the given cable.

After the operation has been chosen the cables are temporarily joined and measurements are carried out to check the resultant coupling and unbalance coefficients of cable A + B.

The agreement of the results of the measurements with the calculated values shows that the operation was chosen correctly.

The method which has been presented for balancing by the transposition method on the basis of the results of measurement of the coupling coefficients (k) and the unbalance coefficients (e) are used for comparatively short sections of cable line (within the limits of a balancing length). In long cable lines, for connecting the balancing lengths together, and especially in loaded cables it is necessary to operate not with the values k and e, but to measure the crosstalk attenuation between the circuits of the balanced quad. At the center of a balanced section of line the cable quads are transposed on the basis of the different operations given above. At the same time the crosstalk attenuation in the cable is checked. The transposition operation is chosen that provides the greatest crosstalk attenuation between the balanced circuits.

In cables used in a four-wire system, the cross-

talk attenuation is measured at the far end, while for a two-wire system the crosstalk attenuation is measured at the near end.

Table 6-22.

Table for Balancing by the Transposition Method. A) Cable; B) sign; C) operation; D) cable; E) sign; F) operation; G) results of checking measurements; H) operation chosen; I) maximum value of sum; J) maximum value of sum; K) note. "+" indicates that the coefficients of A and B have the same sign; "–" indicates that the coefficients of A and B have different signs.

(A) Кабель A	(A) Кабель B	(B) Знак	Оператор (C) ...	Оператор (C) X..	Оператор (C) .X.	Оператор (C) XX.
1	2	3	4	5	6	7
$k_1 = -20$	$k_1 = -30$	−				
		+	50			50
$k_2 = +30$	$k_2 = -20$	−		50		50
		+				
$k_3 = -10$	$k_3 = +20$	−			30	30
		+				
$e_1 = -400$	$e_1 = +300$	−		70		70
		+				
$e_2 = -200$	$e_2 = -100$	−				
		+	30	30		
$e_3 = +300$	$e_3 = -100$	−				
		+				
(I) Максимальное значение суммы			50	70	30	70

(K) Примечание. „+" одинаковые знаки у коэффициентов A и Б „–" разные знаки у коэффициентов A и Б.

Table 6-22
Continued

Кабель D		E Знак	F Оператор				G Результаты контрольных измерений	H Принять оператор
A	B		..X	X.X	.XX	XXX		
8	9	10	11	12	13	14	15	16
$k_1 = -20$	$k_1 = -30$	−					+10	
		+	50			50		
$k_2 = +30$	$k_3 = +20$	−					+10	
		+	50		50			
$k_3 = -10$	$k_2 = -20$	−					−30	
		+	60	60				
$e_1 = -400$	$e_2 = -100$	−					−100	X
		+	50		50			
$e_2 = -200$	$e_1 = +300$	−			50	50	−100	
		+						
$e_3 = +300$	$e_3 = -100$	−	40	40	40	40	+200	
		+						
J Максимальное значение суммы			60	60	50	50		

6-17. BALANCING BY THE METHOD OF ADDITIONAL CAPACITANCES

Capacitive balancing consists of equalizing the capacitive unbalance of a cable with the aid of additional capacitors. This operation is carried out by connecting balancing capacitors of appropriate capacitance between the conductors and between the conductors and ground (the zinc sheathing).

The aim of the technician doing the balancing is to remove the unbalance and to reduce the coefficients k_1, k_2, k_3 and e_1, e_2, e_3 to acceptable values, using the fewest number of additional capacitors.

The method of capacitive balancing is illustrated by the example given below.

Let us assume that measurements of the coupling coefficients of a cable quad yield the following values:

$$k_1 = -30 \ \mu\mu f$$
$$k_2 = +20 \ \mu\mu f$$
$$k_3 = +30 \ \mu\mu f$$

On the basis of the formula given in section 6-6, the equality $k_1 = -30$ µµf shows that the sum of the capacitances $(c_{13} + c_{24})$ is greater than the sum of the capacitances $(c_{14} + c_{23})$ by 30 µµf.

It is evident that by connecting a 30 µµf capacitor

between conductors 1-4 or 2-3, we thereby increase the sum ($c_{14} + c_{23}$) by a value equal to the capacitance of the capacitor which has been connected, and thereby reduce the value of the coefficient k_1 to zero.

However if the capacitor is connected only to one of the conductor pairs (1-4) or (2-3), the values of the coefficients k_2 and k_3 will vary as a result. It may turn out that as a result of removing the interaction between the physical circuits (coefficient k_1) there will be an increase in the interaction between the physical and phantom circuits (k_2 and k_3).

Therefore we do not connect one capacitor for balancing purposes but rather insert capacitors in both pairs of conductors (with lower total capacitance) using a capacitance equal to half the value of the coefficient.

Thus in the example considered it is necessary to connect two capacitors of 15 μμf each: one of them between conductors 1-4, the other between conductors 2-3. This will ensure that the equality $k_1 = 0$ has no effect upon the values of the coefficients k_2 and k_3.

In like manner capacitors are chosen to balance k_2 and k_3. Here k_2 is made to equal zero by connecting capacitors of 10 μμf each into conductors 2-3 and 2-4, while k_3 is set to zero by connecting 15 μμf capacitors

in conductors 1-4 and 2-4.

The results of balancing the cable quad under consideration are given in Table 6-23, where in the column headed "sum" we find the values of the capacitors which had to be connected between the appropriate conductors of the cable quad in order to remove interaction between the circuits.

Table 6-23.

The Principle of Capacitive Balancing of a Cable Quad. A) Coupling coefficients; B) values of the balancing capacitors required between the conductors; C) checking measurements; D) sum; E) calculated least value; F) capacitors connected.

Коэффициенты связи	Величины симметрирующих конденсаторов, которые необходимо включить между жилами				Контрольные измерения
	1—3	1—4	2—3	2—4	
$k_1 = -30$	15	—	—	15	
$k_2 = +20$	—	—	10	10	
$k_3 = +30$	—	15	—	15	
Сумма	15	15	10	40	
Вычитаемая наименьшая величина	10	10	10	10	
Включаемые конденсаторы	5	5	0	30	

It is evident that the capacitive balance of a quad is not upset when the values of all four capacitors are decreased by the same amount. Accordingly subtracting the value of the smallest capacitor (which was 10 µµf in the example considered), we obtain the final values of the balancing capacitors which must be inserted in the cable quad.

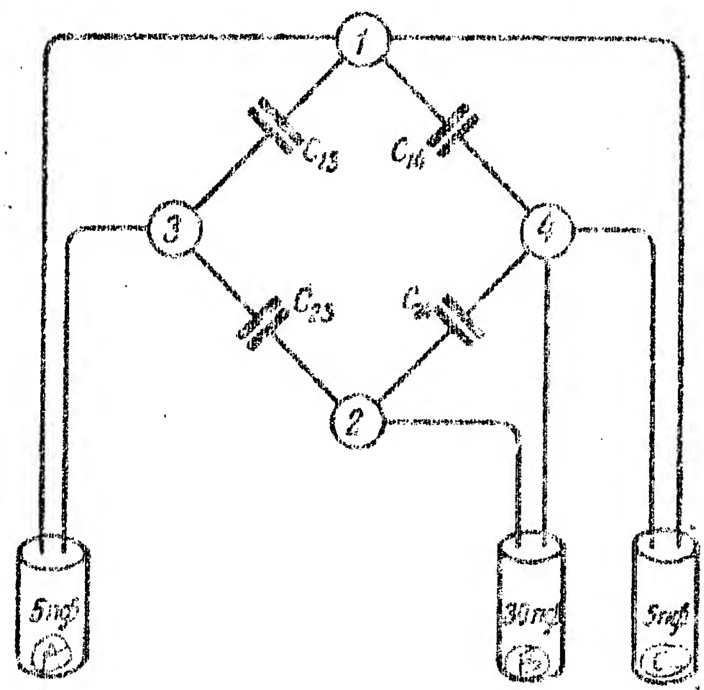


Fig. 6-51. The connection of balancing capacitors into a cable quad. A) 5 μμf; B) 30 μμf; C) 5 μμf.

In like manner the quad is balanced with respect to ground and the unbalance coefficients e_1, e_2 and e_3 are compensated.

Figure 6-51 illustrates capacitance balancing of the quad which has been considered.

Table 6-24.

Balancing by the Method of Additional Capacitors. A) Balancing with respect to ground (zinc sheathing); B) balancing between conductors of the cable; C) results of measurements of unbalance coefficients; D) fraction of the measured values which must be inserted in empty box; E) capacitors connected between conductors and ground; F) results of measurements of coefficients of coupling; G) fraction of measured value which must be inserted in empty box; H) capacitors connected between conductors of the cable; I) coefficient; J) sign; K) value; L) coefficient; M) sign; N) value; O) sum; P) sum; Q) computed least value; R) computed least value; S) capacitance; T) capacitance.

A Симметрирование по отношению к земле (свинцовой оболочке)

C Результаты измерения коэффициентов асимметрии			D Часть измеренного значения, которую необходимо включить в чистую клетку	E Конденсаторы, включаемые между жилами и землей			
H Коэффициент	J Знак	K Величина		1—0	2—0	3—0	4—0
1	2	3	4	5	6	7	8
e_1	—		1, 1/2, 1/2				
	+	200	1, 1/2, 1/2		200	100	100
e_2	—	100	1/2, 1/2, 1	50	50	100	
	+		1/2, 1/2, 1				
e_3	—	300	1/2, 1/2	150	150		
	+		1/2, 1/2				
Сумма O				200	400	200	100
Q Вычитается наименьшая величина				100	100	100	100
S Емкость конденсатора				100	300	100	0

B Симметрирование между жилами кабеля

E Результаты измерения коэффициентов связи			G Часть измеренного значения, которую необходимо включить в чистую клетку	H Конденсаторы, включаемые между жилами кабеля			
L Коэффициент	M Знак	N Величина		1—3	1—4	2—3	2—4
9	10	11	12	13	14	15	16
k_1	—		$1/_2$, $1/_2$				
	+	20	$1/_2$, $1/_2$		10	10	
k_2	—		$1/_2$, $1/_2$				
	+	40	$1/_2$, $1/_2$			20	20
k_3	—	20	$1/_2$, $1/_2$	10		10	
	+		$1/_2$, $1/_2$				
P Сумма				10	10	40	20
R Вычитается наименьшая величина				10	10	10	10
T Емкость конденсатора				0	0	30	10

Balancing capacitors are cylindrical in shape and are manufactured in a wide variety of capacitances, varying by 5-10 μμf. Capacitors are connected to the conductors of the cable at the transposition point of the shipping length sections.

A general view of a capacitor joint is shown in Fig. 6-52.

In practice capacitive balancing is carried out with the aid of special, previously prepared tables which considerably ease the task of balancing.

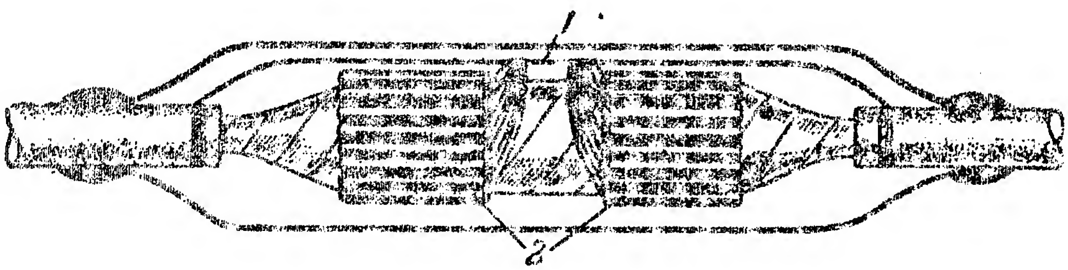


Fig. 6-52. Capacitor joint. 1) Conductor for connecting capacitors to zinc sheathing; 2) balancing capacitors.

Table 6-24 gives an example of capacitive balancing with the aid of a special table.

The table consists of two portions including the balancing of quads with respect to earth (e_1, e_2, e_3) and balancing between the conductors of the quads (k_1, k_2, k_3).

The results of measuring the unbalance and coup-

ling coefficients are given, respectively, in columns 1-3 and 9-11. The corresponding factions of the measured values are inserted in the empty boxes opposite the sign of the measured value.

The added absolute values are found in the vertical column "sum." The least sum is chosen of those obtained; thus the values are obtained for the capacitances which must be connected in the cable quad in order to compensate the capacitive coupling and unbalance.

Capacitive balancing to remove capacitive unbalance of circuits within cable quads has been described above.

Capacitive balancing of adjacent quads is carried out by a similar method. In this case the values of the coefficients k_{9-12} and k_{4-8} are measured (where phantom circuits are used) in order to choose the capacitors which must be connected between the circuits of the two balanced quads to decrease the coupling coefficients to acceptable values.

In order to guard against interference between circuits of adjacent quads, in addition to balancing with capacitors, mixing of the quads is commonly used. This consists in changing the position of the quads along the extent of a cable line, with the quads periodically get-

ting further apart and drawing together again.

Thus for example if in the first factory length of cable, quad No. 1 and quad No. 2 are located together, then in the second factory length, quad No. 2 is separated from the first quad by one quad and is connected not with the second quad but with the third quad and so forth. With such quad mixing the relationship of any two quads with respect to position recurs rather seldom (for example with 15 quads in a layer -- only in every seven factory lengths).

Quad mixing is quite effective and commonly used, especially in cables having a large number of quads.

6-18. CONCENTRATED BALANCING

Concentrated or lumped balancing consists of removing mutually interfering effects between cable circuits by compensating couplings at one or two points of the cable line. The length of cable for which concentrated balancing is carried out is equal, as a rule, to the repeatered section, i.e., amounts to 25-120 km depending upon the type of cable.

Coupling compensation is accomplished by connecting special equalization circuits, called anti-coupling elements, into the line between the balanced circuits.

Concentrated balancing is chiefly used to protect

high-frequency symmetrical cables against interference.

The advantage of concentrated balancing lies in the great economy of a method which permits balancing a very long cable at only 1-2 points, since with normal balancing measures the coupling must be compensated at each shipping length section.

The concentrated-balancing method is based upon the fact that electromagnetic coupling is complex in nature and its action is expressed by the resultant vector, and it is certainly possible to select anti-coupling elements which will compensate for the effect.

With the aid of anti-coupling elements, an artificial current I_k is set up in the circuit experiencing the interference; this current is equal in value and opposite in sign to the noise current, $I_k = -I_n$. As a result the currents cancel and the noise in the cable is considerably reduced.

The anti-coupling element must be equal in absolute value to the resultant vector of the natural coupling, and have a phase (angle) of opposite sign.

Thus, for example, if as a result of measurements it has been established that the vector of the natural coupling in the cable is:

$$K = 50\ e^{+j120^\circ},$$

then the introduced compensating element must equal:

$$K_{\kappa} = 50\ e^{+j(120^\circ + 180^\circ)} = 50\ e^{-j60^\circ}$$

Fig. 6-53 gives the basic circuit for connecting an anti-coupling element between two balanced circuits; it shows that if the natural coupling in the cable leads to the appearance of a noise current I_n, then the current which is occasioned by the connection of the anti-coupling element, $\dot{I}_k$, is equal in value to the noise current and is opposite in direction

$$\dot{I}_{\kappa} = -\dot{I}_{n}.$$

As a result these currents cancel out, decreasing the interaction of the balanced circuits.

It is necessary to break down the peculiarities in using the method of concentrated balancing on the basis of the various systems in which cables must be used; the reason for this is that the effectiveness of the method in protecting against interfering effects will be different at the near and far ends of the cable.

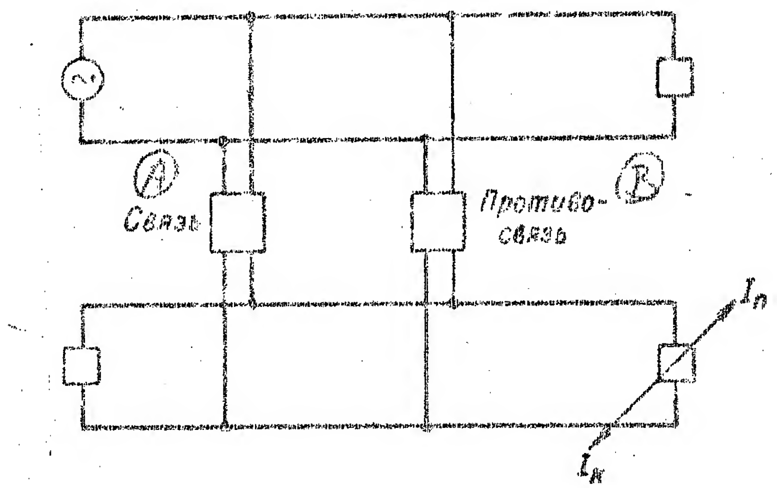


Fig. 6-53. Circuit for connecting a compensating anti-coupling element. A) Coupling; B) anti-coupling.

As a rule the method of concentrated balancing permits the elimination of interaction only at the far end of a cable. The reason for this is that in order to use concentrated balancing for interaction at the near end of a cable it is necessary to know accurately the local effect of the resultant vector of the natural coupling and to connect the anti-coupling element at exactly that point

It is almost impossible to establish the effect at this point.

There is no such requirement for concentrated balancing at the far end of a cable, and the anti-coupling element, even if connected in the middle of the line, provides the necessary protection of the equipment against interference.

Figure 6-54 shows the paths of the interference and compensating currents at the near and far end of the cable. At a distance l_n the vector of the natural coupling acts, while the compensating element has been shown to be connected at a distance l_k. It is clear from Fig. 6-54 that in the case of interaction at the far end of the cable the path of the noise current due to the natural coupling, I_n, equals the path of the compensating current flowing through the anticoupling element, I_k. This equality of the current paths for I_n and I_k holds wherever the anti-coupling element is connected in the cable line. Since the currents I_n and I_k are opposite in phase, $\dot{I}_k = -\dot{I}_n$, and the interference effect in the cable at the far end is sharply reduced.

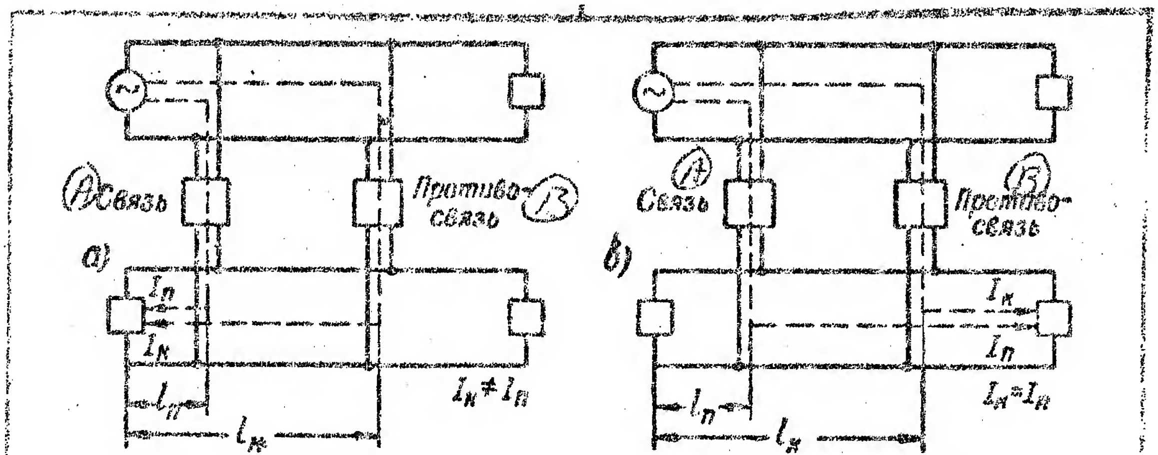


Fig. 6-54. Basic difference in compensating coupling at the near and far end of a cable. a) Interaction at the near end; b) interaction at the far end. A) Coupling; B) anti-coupling.

At the near end of the cable the strength and phase of the anticoupling current depend upon the point at which the compensating anticoupling element is connected. As may be seen from Fig. 6-54, the paths of the currents $\dot{I}_n$ and $\dot{I}_k$ are different. If $l_k > l_n$, then the current path $\dot{I}_k$ is greater than the current path for $\dot{I}_n$, and consequently the compensating current, suffering greater attenuation, is considerably weaker at the near end than is the noise current I_n.

Compensation does not even take place for $l_k < l_n$, the case where the current I_k is not so strongly attenuated as the current I_n. Here their phases will differ but the shift will not equal 180 degrees. As a result a difference

current (noise) will be created in the equipment associated with the circuit subject to the effect. Consequently concentrated balancing of the effect at the near end does not yield a satisfactory effect. Owing to this fact the concentrated balancing method is chiefly used for cable trunks used in a two cable system, where interaction between circuits at the far end of the circuit plays the chief role.

The following information is necessary in the design and construction of compensating anticoupling circuits.

The electromagnetic coupling coefficients may have arbitrary magnitude and a phase lying within limits of 0 to 360 degrees.

The coupling vector may lie in any of the four quadrants of a rectangular coordinate system. In accordance with this the circuits and elements of the anticoupling circuit must be capable of providing an anticoupling vector of any amplitude or phase.

The anticoupling circuits may be formed from resistors and capacitances R and C, or from resistors and inductances R and L.

In practice the most common type of concentrated balancing for high frequency cables utilizes RC circuits; the circuits, as a rule, take the form of series-connected high resistances and capacitors. Parallel-connected RC cir-

cuits are not used, since they have a harmful effect on the insulation resistance and other parameters of the cable itself.

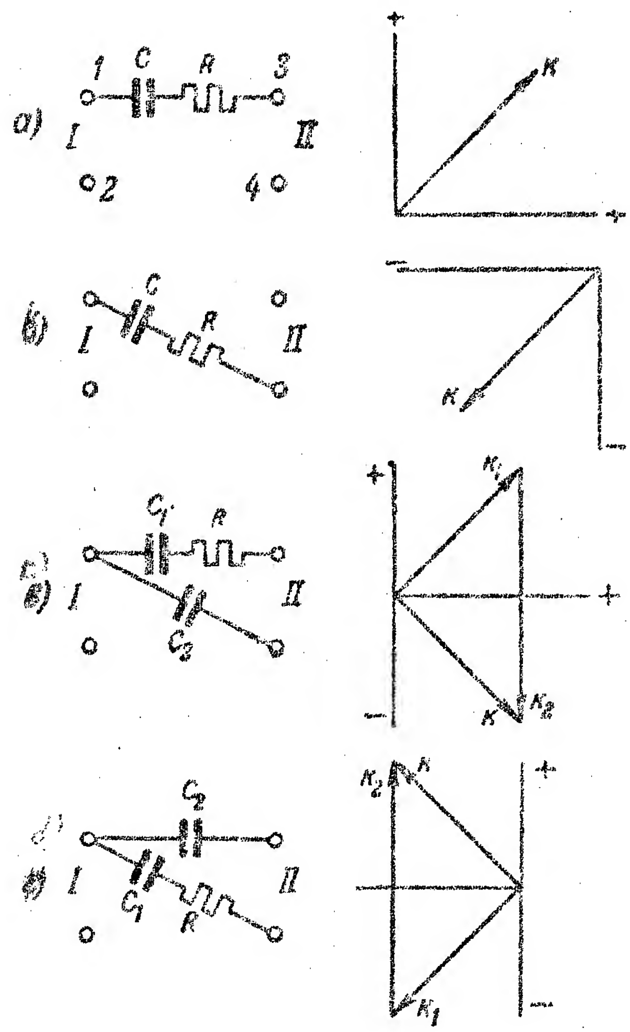


Fig. 6-55. Circuits of anticoupling elements. a) First quadrant; b) third quadrant; c) fourth quadrant; d) second quadrant.

The phase of the anticoupling vector is regulated by choosing values of R and C and connecting them in the appropriate conductors of the balanced quad.

Fig. 6-55 shows anticoupling RC circuits for various positions of the anticoupling vector. Conductors 1-2 belong to the first circuit, conductors 3-4 to the second.

Series connection of the resistance R and the capacitance C across conductors 1-3 gives a coupling vector varying within the limits of the first quadrant alone.

The third quadrant may be covered by reversing the sign of the anticoupling vector, which is done by connecting the antinoise element across conductors 1-4.

If the coupling vector must be located in the second or fourth quadrant, the purely capacitive elements must be connected in parallel.

For this reason the vector K_2 must be added to the original vector K_1, and the resultant anticoupling vector K will fall in the appropriate position in the second or fourth quadrant.

Figure 6-55c shows the circuit of an anticoupling element where the vector K is located in the fourth quadrant, while Fig. 6-55d shows the vector K located in the second quadrant.

Another method of concentrated balancing utilizes

anticoupling elements based on special coils consisting of resistances R and inductances L.

The principles by which such coils act is shown on Fig. 6-56. The primary winding of the coil consists of two

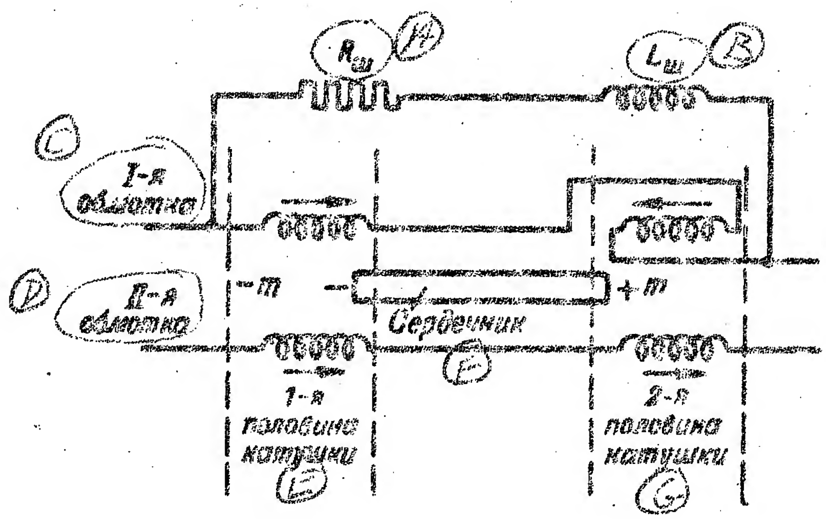


Fig. 6-56. The principle of action of compensating inductance coils. A) R_{sh}; B) L_{sh}; C) primary; D) secondary; E) first half of the coil; F) core; G) second half of the coil.

series-connected halves with turns wound in opposite directions. The secondary winding consists of the same two halves with the turns wound in the same direction. As a result the inductive coupling between the primary and secondary windings will be zero since the coupling contains

two components which are equal in magnitude but opposite in sign. There is a core in the coil consisting of two rigidly fastened but mutually insulated portions. When this core is located exactly in the center (zero position), the coupling between the windings will remain equal to zero.

If the core is shifted to one side or another, then coupling of one side or another begins to predominate. In order to give the inductance of the coil a complex character, corresponding in frequency variation to the real and imaginary component of the electromagnetic coupling of the cable, and inductive-resistive shunt is connected across one of the windings of the coil, with the appropriately chosen resistance R_{sh} and inductance L_{sh}.

There are also inductive coils in which the complex character of the anticoupling vector has been obtained not by using an inductive-resistive shunt, but by using a special moving wheel. The circuit of such a coil is given in Fig. 6-57. Here the coil is located in a screen, above which a movable metal wheel is located. There is also a movable core within the coil. By moving the core it is possible to vary the imaginary component of coupling, as well as its phase, and by moving the wheel along the screen from one end to the other it is possible to regulate the real component of the coupling within specific limits.

It should be kept in mind that if the disturbed circuit II and the disturbing circuit I are interchanged, the values of noise and the corresponding results of measurement of the coupling coefficients (measuring I-II, and then II-I) may differ. This is explained, primarily by the difference in the propagation constants of the first, γ_1, and second, γ_2, circuits, as well as by indirect interaction through a third circuit, and nonuniformity in the circuits themselves.

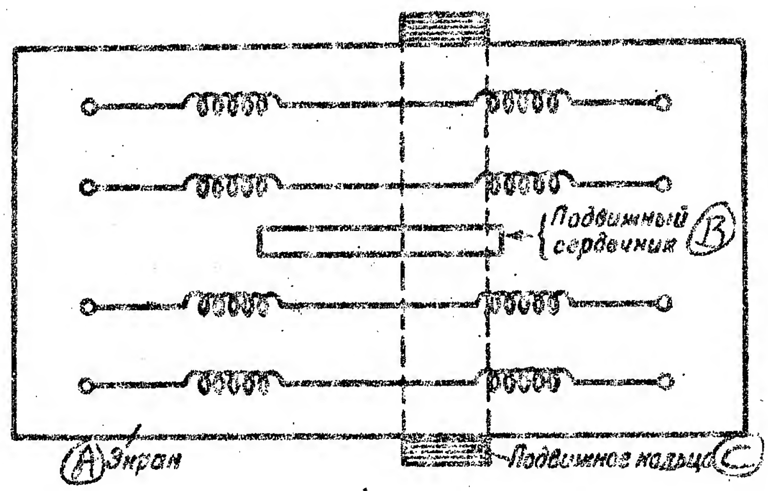


Fig. 6-57. Circuit of a compensating coil with a moving wheel. A) Screen; B) movable core; C) movable wheel.

On account of this compensation of interaction at one point does not give satisfactory results, owing to the difference of coupling vectors in the two cases of inter-

action and the necessity of compensating for them by using different anticoupling circuits.

This factor shows up especially strongly when compensating couplings over a wide frequency band.

It has been shown both theoretically and experimentally that if concentrated balancing is carried out at two points of the cable line, it will provide satisfactory compensation of coupling. It has been found in practice, that especially satisfactory results are obtained when balancing is used at two points A and B located at distance $A = {}^{1}/_{3}\underline{1}$ and $B = {}^{2}/_{3}\underline{1}$ where $\underline{1}$ is the length of a repeatered section of the cable line.

Thus for example if a repeatered section of a high frequency link equals 30 km, the concentrated balancing elements are connected at the points A = 10 km and B = 20 km.

There are long line trunk cables in which the compensating elements for coupling are installed not in the line but at the repeater points and the terminals on special frames. In general, in these cases, the concentrated balancing is carried out with the aid of inductance coils (R and L circuits).

Concentrated balancing is carried out on the basis of the results of measurements of the resultant vectors of

electromagnetic coupling at the highest frequency transmitted. However, it is necessary to check the coupling at other frequencies as well, in an effort to provide satisfactory protection against interference over the entire range of frequencies used in practice.

In addition to the insertion of lumbed equalization elements to compensate for coupling, concentrated balancing by means of isolated transposition is also widely used.

Concentrated transposition, as a rule, is carried out at one of two points in a length of cable equal to a repeatered section. In the majority of cases the circuit is transposed at the same points at which compensating anticoupling elements are connected, i.e., at the points A = 1/3 and B = 21/3.

Balancing by the method of concentrated transposition is carried out on the basis of the results of measurements of the cross-talk attenuation between circuits I and II, normally at the far end of the cable line. The cross-talk attenuation is measured for different combinations of the transposition operations at points A and B, and the cross-talk-attenuation measurements are carried out for two interaction combinations: from the first circuit to the second — $B_{I/II}$ and from the second circuit to the first — $B_{II/I}$.

Table 6-25 gives the possible variations for the combinations of transposition operations at points A and B.

Phantom circuits are not used in high-frequency communications cables; thus the operations associated with the indexes for phantom-circuit transpositions are not shown in Table 6-25.

On the basis of the results of measurements of the cross-talk attenuations $B_{I/II}$ and $B_{II/I}$, at points A and B, the two operations are chosen which yield the largest value of cross-talk attenuation.

All measurements and the choice of operations are carried out at the highest transmitted frequency, while at the same time the cross-talk attenuation is checked at intermediate frequencies.

Cable balancing normally begins with the use of the concentrated transposition method; later, quads which do not satisfy the standards are additionally balanced by the method of connecting compensating anticoupling elements at widely separated points.

Table 6-25.

A № сочетаний	B Оператор скрещивания в точке:		C Результаты измерения переходного затухания (непер)	
	А	Б	$B_{I/II}$	$B_{II/I}$
1	..	..		
2	X.	..		
3	.X	..		
4	XX	..		
5	..	X.		
6	X.	X.		
7	.X	X.		
8	XX	X.		
9	..	.X		
10	X.	.X		
11	.X	.X		
12	XX	.X		
13	..	XX		
14	X.	XX		
15	.X	XX		
16	XX	XX		

Various Combinations of Operations for Concentrated Transposition. A) Combination No.; B) operation of transposition at point:; C) results of measuring cross-talk attenuation (nepers).

6-19. EQUALIZATION OF EFFECTIVE CAPACITANCES AND PURE RESISTANCES

In order to improve the uniformity of cable circuits, ~~and correspondingly to decrease the~~ waveshaping ~~ability of~~

the input impedance, and to increase the cross-talk attenuation, during the assembly of cable lines the effective capacitances are equalized and the resistive unbalances of the conductors are removed.

This is especially important in coil-loaded cables and circuits, used in carrier-frequency multiplexing.

The permissible deviations for the effective capacitances and resistive unbalance of conductors is given in section 6-8.

The effective capacitances are equalized in two stages. First factory lengths of cable are divided into 4-8 groups so that the average working capacitance of any group differs from the average working capacitance of all the cables of a repeatered section by not more than ±2%.

Thus for example if the average working capacitance of a repeatered section of cable is 0.030 μf, then the effective capacitance of the first group should be not less than 0.0294 μf, and the effective capacitance of the next group, not less than 0.0306 μf. The capacitances of the remaining groups should lie, in steps, within the limits of 0.0294-0.0306 μf.

The cable should be laid in such fashion that the factory lengths of cable of one group are placed next to those of the adjacent group on the basis of increasing or

decreasing numbers of the group.

The second stage in equalizing the capacitances consists in the choice of quads in joints when the cable is assembled. The choice of quads is carried out so as to decrease the deviation of the effective capacitance of the circuit from the average effective capacitance. In order to do this, when the factory lengths of cable are assembled, the quads with the greatest positive capacitance deviation are singled out and connected to the quads having the greatest negative capacitance deviation.

The selection of quads on the basis of the effectiv capacitances is chiefly carried out at the capacitor joints and the coil-loading boxes.

If the resistive unbalance of the conductors exceeds the permissible standard, it too is equalized. In order to do this, high-resistance conductor is soldered to the conductors having less resistance. For this purpose, as a rule, insulated constantan wire is used, 0.8 mm in diameter.

It should be noted that the input impedance of a cable circuit is basically determined by the uniformity of the first 10-20 cable lengths from both ends of a repeatered section; therefore they should be balanced and equalized with especial care.

CHAPTER SEVEN

INTERACTION IN COAXIAL CABLES

7-1. THE NATURE OF INTERACTION IN COAXIAL CIRCUITS

When current is passed through symmetric circuits, there appear about them external fields: electrical (E_r and E_φ) and magnetic (H_r and H_φ) (Fig. 7-1).

If any circuit II falls within the sphere of action of the electromagnetic field of a circuit I, then there will be induced a current in circuit II, which will take the form of interference with the basic transmission being sent over circuit II.

The electromagnetic interaction between symmetrical circuits has come to be expressed with the aid of the capacitive and inductive coupling coefficients or in terms of the cross-talk attenuation.

It has been shown above that coaxial circuits have no external lateral electromagnetic fields of the E_r, E_φ type or of the H_r, H_φ type.

The radial electrical E_r and tangential magnetic H_φ fields of a coaxial circuit are short circuited within the cable between the inner and outer conductors; the E_φ and H_r fields are absent owing to the axial symmetry of the cable.

Thus a coaxial circuit II, located next to a coaxial circuit I, through which energy is being transmitted, will not experience the influence of electromagnetic fields in the radial and tangential directions.

For this reason it would appear that these circuits should not experience the effect of mutual interference, unlike symmetric circuits. In actuality however the opposite is true, since adjacent coaxial circuits do affect each other and are susceptible to extraneous sources of noise (radio stations, electrical power transmission lines, etc.).

Coaxial cables are liable to mutual and external interference owing to the longitudinal components of the electrical field directed along the axis of the coaxial cable, E_z.

Of course in symmetric circuits there is also a voltage drop along the conductors, and a longitudinal electrical field E_z, but its effect is considerably weaker than that of the radial and tangential fields, and thus when the processes of interactions in these circuits are considered, it may be neglected.

In coaxial cables, where there are no external radial or tangential fields, interference is determined precisely by the longitudinal field.

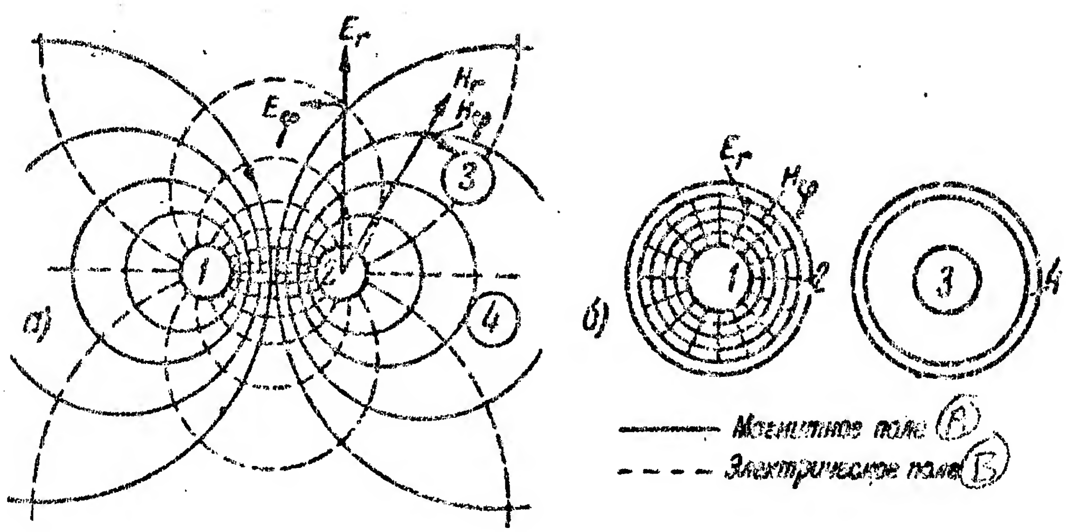


Fig. 7-1. Electromagnetic field. a) Symmetric construction; b) coaxial construction. A) Magnetic field; B) electrical field.

The interaction of two coaxial circuits I and II occurs by way of a third intermediate circuit, formed by the outer conductors of the circuits.

As can be seen from Fig. 7-2 three circuits participate in the interaction of coaxial circuits:

I — the disturbing circuit;

II — the disturbed circuit;

III — the intermediate circuit consisting of the outer conductors of I and II.

The physical interaction between two coaxial cables may be represented in the following manner.

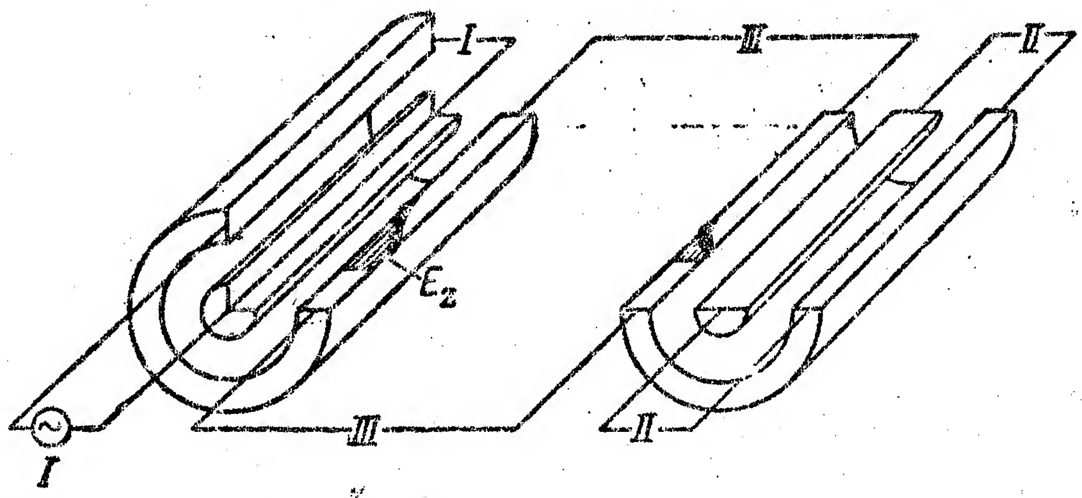


Fig. 7-2. Interaction circuit in coaxial cables. I) Disturbing circuit; II) disturbed circuit; III) intermediate circuit.

The current flowing through the outer conductor of the disturbing coaxial cable I sets up a voltage drop on its external surface; in connection with this the longitudinal electric-field component E_z acts. This sets up a current on the external surface of the outer conductor of the disturbed cable II. Thus, an intermediate current loop is formed by the two outer conductors of the cable; an emf equal to the E_z on the external surface of the outer conductor of the disturbing cable acts in this circuit.

The current flowing along the outer conductor of the disturbed cable creates a voltage drop which sets up interference in the circuit.

Thus the mechanism of interaction of coaxial cables is as follows: the interfering circuit I creates a voltage and current in a circuit III, which in turn sets up an interfering circuit with respect to circuit II, inducing noise currents in it.

The intensity of the intercircuit interference is determined by the strength of the longitudinal component of the electrical field, E_z, on the external surface of the outer conductor of the interfering coaxial circuit. The greater the magnitude of E_z, the greater the voltage and current in the intermediate circuit III, and correspondingly the noise current in the disturbed circuit.

The frequency dependence of coaxial-cable interference is in principle different than that of symmetric circuits. While in the latter the intercircuit interference increases with frequency and the resistance to external noise drops, the opposite is true of coaxial cables. In coaxial cable transmission, the least noise resistant frequency band lies in the 50-60 kc region, as illustrated in Fig. 7-3.

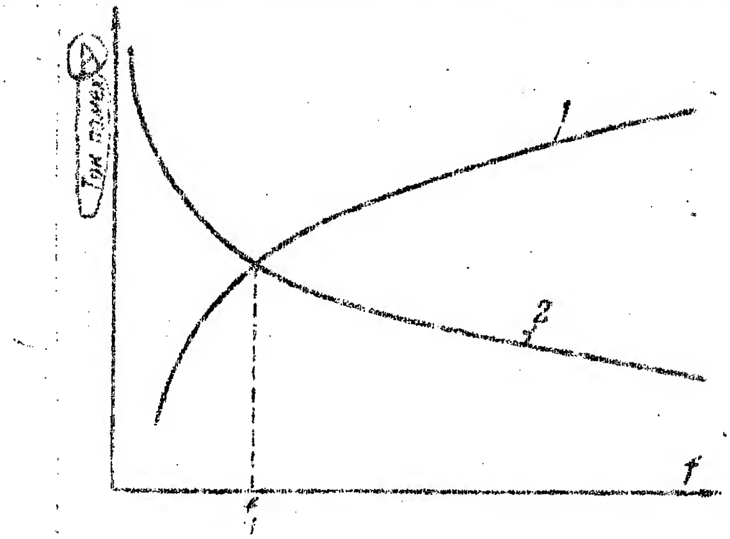


Fig. 7-3. The frequency-dependence of interference in coaxial and symmetric cables. 1) Symmetric; 2) coaxial cable. A) Noise current.

This is explained by the fact that, owing to the proximity effect in coaxial cables, the current density in the outer conductor increases from outside to inside surface, and as the frequency increases the current has a tendency to concentrate on the inner surface of the outer conductor, while on the outer surface the density decreases. Thus as the frequency increases the field strength E_z decreases on the outer surface of the outer conductor, and the self-shielding effect of a coaxial cable increases.

At very high frequencies, where all of the current is concentrated within the coaxial cable, the field strength

E_z beyond the cable tends to zero, the screening effect reaches a maximum, and the interaction of the circuits theoretically vanishes.

The interaction of coaxial circuits also depends upon the structure of the outer conductors, their physical arrangement, and the material from which they are manufactured. In particular the thicker the outer conductor the less the interaction.

Steel has better screening properties than copper, and thus for protection from interference, chiefly in the high frequency region, the surface of the copper outer conductor coaxial cable is covered with two spiral layers of steel tape.

Just as in symmetric circuits, the interaction in coaxial circuits is expressed and standardized in terms of the cross-talk attenuation at the near end, B_0, and at the far end, B_1 of the cable.

In considering questions connected with interference in coaxial cables, another parameter is used called the coupling impedance or the cross-talk impedance, Z_{12}.

This parameter Z_{12} is formally similar to the electromagnetic coupling coefficient in symmetric circuits. It intrinsically corresponds to the resistive coupling

coefficient r.

In addition to the interaction of coaxial circuits it is also necessary to consider their liability to interference from powerful radio stations.

The electromagnetic field radiated by the antenna of a radio transmitter is propagated through the air, penetrates into the ground, falls upon the sheath of the cable and the coaxial circuit within. The interference effect is due to the horizontal component of the electrical field, E_h (Fig. 7-4).

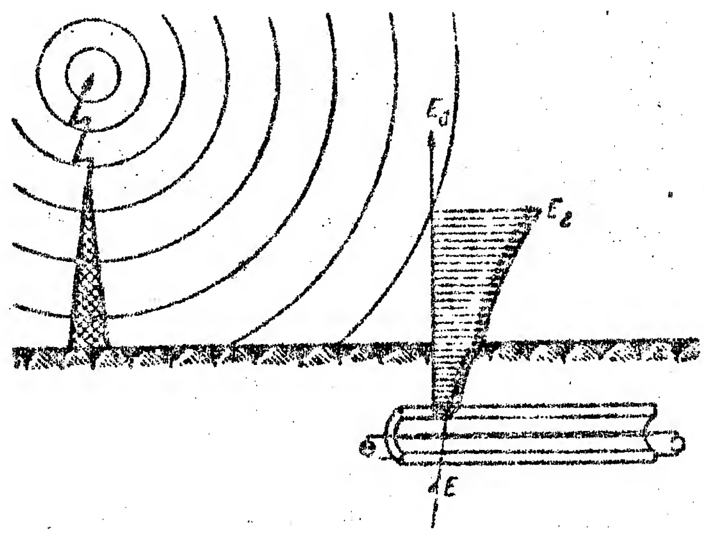


Fig. 7-4. The effect of radio stations on communication along a buried coaxial cable. E_v — Vertical component of the electrical field; E_h — horizontal component of the electrical field. A) E_v; B) E_h.

It has been established experimentally that in unshielded coaxial cables, radio interference is felt up to 40 kc, while in shielded cables it is felt up to 15 kc. Underground coaxial cable trunks, used for frequencies of 60 kc and above, as a rule, are free of radio station interference.

7-2. COUPLING IMPEDANCE

The coupling impedance Z_{12} takes the form of a relation of the voltage excited on the external surface of the outer conductor of a coaxial cable, $\dot{U}_c$, to the current flowing in the coaxial circuit, $\dot{I}$. Keeping in mind that the voltage $\dot{U}_c$ corresponds to the longitudinal component of the electric field on this surface of the conductor, we may write:

$$Z_{12} = \frac{\dot{U}_c}{\dot{I}} = \frac{\dot{E}_z^c}{\dot{I}}.$$

As can be seen from Fig. 7-5, when a current is passed through a coaxial circuit, a voltage drop is set up on the outer conductor, and a longitudinal electric-field component $\ddot{E}_z$ comes into play: this causes interference between the circuits.

A relationship of the magnitude of E_z to the cur-

rent in the circuit permits qualitative evaluation of the coupling impedance. The greater Z_{12}, the greater E_z at the external surface of the outer conductor of the coaxial cable and beyond it, and the greater the interference of the given circuit with others.

The coupling impedance Z_{12} determines the strength of the field about the coaxial cable, characterizes the amount of energy transmitted along the interfering coaxial circuit I, transferred to the intermediate circuit III, and thence to the disturbed circuit II.

The longitudinal electric-field component on the external surface of the outer conductor of a coaxial cable, E_z^c, may be represented as the product of the current density at this surface J^c and the resistivity of the metal, ρ.

$$\dot{E}_z^c = \dot{J}^c \rho.$$

Since $Z_{12} = \dot{E}_z^c/\dot{I}$, we obtain $Z_{12} = \rho(\dot{J}^c/\dot{I})$, from which it follows that the coupling impedance Z_{12} is directly proportional to the current density J^c at the outer surface of the coaxial cable.

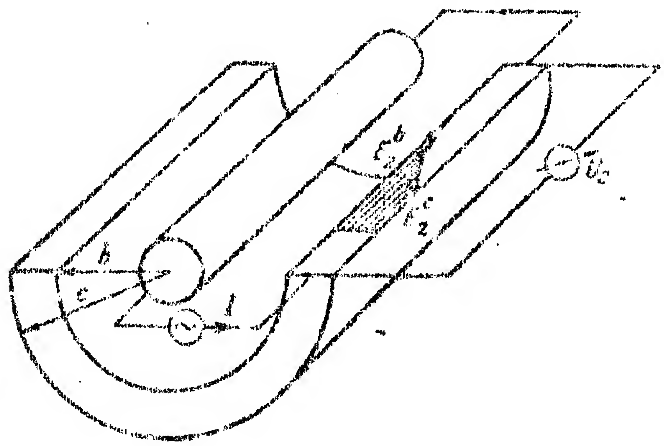


Fig. 7-5. Diagram for the discussion of
the coupling impedance of a coaxial cable.

Fig. 7-6 shows the pattern of distribution of current density along the outer conductor of a coaxial cable at various frequencies. The current-density vector $\vec{J}^c$ characterizes the vector of the coupling impedance Z_{12}. The greater the current density at the surface of the outer conductor, the greater the coupling impedance and the greater the longitudinal component of the field strength beyond the cable.

It was previously shown that as the frequency increases an ever sharper redistribution of current takes place, as the current density on the external surface of the outer conductor of the cable decreases. A similar law holds for the coupling impedance as well. The greater the frequency the less its effect. The coupling impedance z_{13}

is at a maximum for DC, and is numerically equal to the DC resistance of the outer conductor of the cable, R_0

$$Z_{12} = \rho \frac{J^c}{I_0} = \frac{U^c}{I_0} = \frac{E_z^c}{I_0} = R_0.$$

This is the case in which the greatest amount of the energy transmitted over circuit I is transferred to circuit II in the form of noise.

As the frequency of the transmitted current increases, Z_{12} decreases and the interference between the coaxial circuits decreases correspondingly.

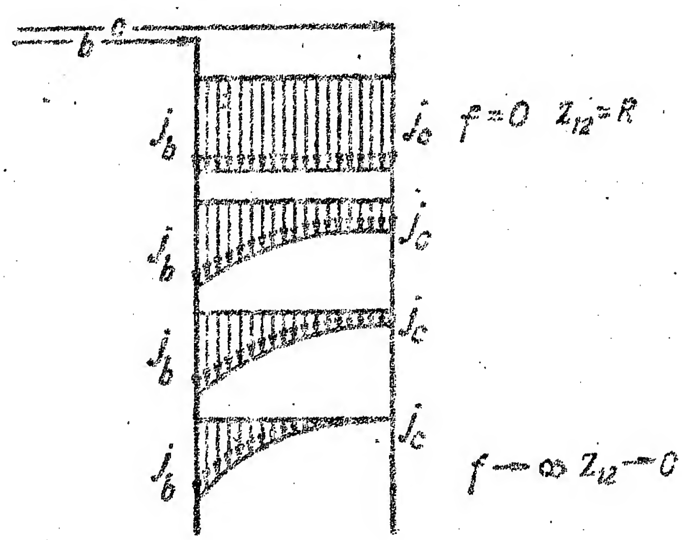


Fig. 7-6. Distribution of current density along the outer conductor of a coaxial cable at different frequencies.

The entire discussion presented above refers to

an interfering coaxial cable (the energy source lies within the circuit). However it also applies to the case where the disturbed circuit is subjected to the influence of an energy source located outside the circuit. However in this case, the greatest current density will be found not at the inner but at the outer surface of the outside conductor of the coaxial cable (Fig. 7-7).

Two coupling impedances are involved in the transfer of energy from the first coaxial circuit to the second: Z'_{12} — for the disturbing circuit, and Z''_{12} for the disturbed circuit.

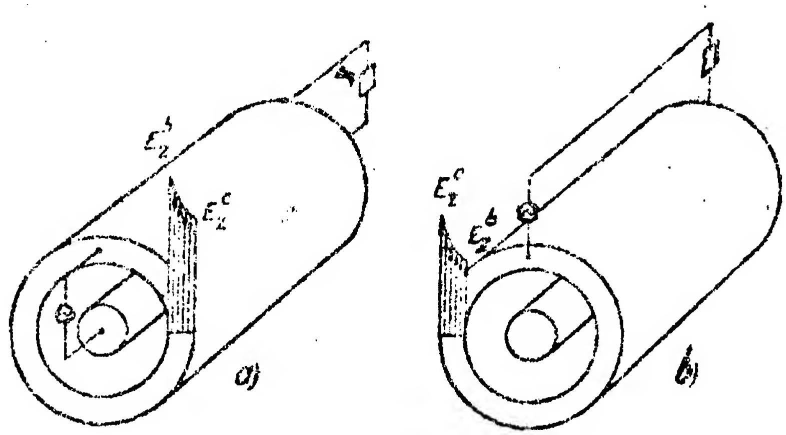


Fig. 7-7. The distribution of E_z and the corresponding current density along the outer conductor of a coaxial cable. a) The source of energy lies within the circuit (disturbing circuit); b) the source of energy lies outside the circuit (disturbed circuit).

Fig. 7-8 gives the diagram for determining the coupling impedances of two coaxial cables.

In order to compute the magnitude of the coupling impedances for a coaxial cable within the frequency band used in practice the following formula is used:

$$Z_{12} = Z_{вс} = \frac{\sigma}{\gamma_1} \frac{1}{2\pi\sqrt{bc}} \cdot \frac{1}{\sinh \sigma t} \ [ohm/см], \qquad (7\text{-}1)$$

where $\sigma = \sqrt{j\omega\mu_1\gamma_1}$ is the eddy-current coefficient;

b and c are the inner and outer radii of the outer conductor of the coaxial cable, in cm;

t is the thickness of the outer conductor, in cm.

In practical units: $\mu_1 = 4\pi\mu \cdot 10^{-9}$, $\gamma_1 = \gamma \cdot 10^4$.

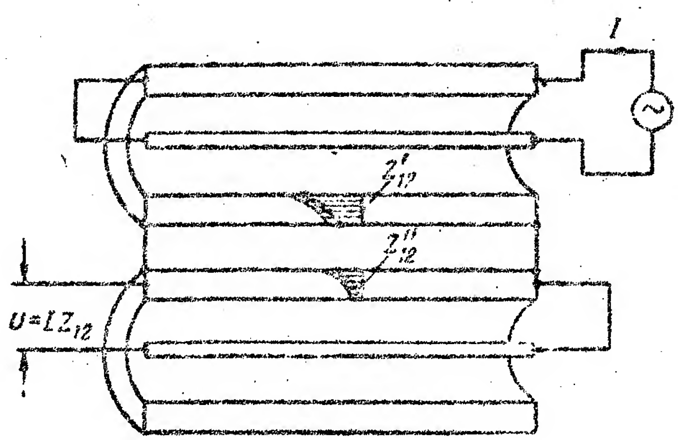


Fig. 7-8. Determination of coupling impedances of two coaxial cables $z_{12} = U/I$.
z_{12} is the coupling impedance (resultant);

U is the voltage in the II circuit; I is the current in circuit I.

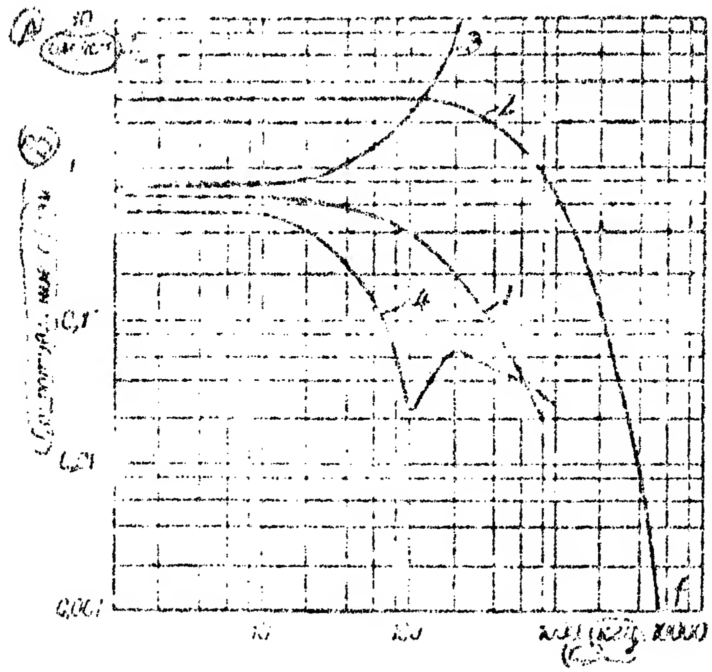


Fig. 7-9. Coupling impedance of various types of outer conductors for coaxial cables. 1) Closed copper shell; 2) closed lead shell; 3) shell of copper tape; 4) laminated shell. A) ohms/km; B) coupling impedance; C) kc.

In the very high frequency region, where $\sigma t \geq 3$, the quantity

$$Z_{12} = \frac{\sigma}{\gamma_1} \cdot \frac{1}{\pi b} e^{-\sigma t} \quad [ohm/cm]. \qquad (7-2)$$

From the formulas which have been given, it follows that the coupling impedance decreases sharply as the

frequency increases or the thickness of the shell becomes greater.

Fig. 7-9 shows the frequency dependence of Z_{12} for copper and lead closed shells, as well as for a laminated shell, and a shell made of copper tape. It is clear from the graph that copper presents better protection against noise than does lead, while the best result is given by the laminated shell. While for the closed shells Z_{12} decreases as the frequency rises, Z_{12} increases with the tape shell and its ability to withstand interference drops. This is explained by the presence in the shells which have been wound of longitudinal external magnetic fields.

In the type 2.6/9.4 trunk cable, a bimetallic copper-steel shell is used; in this case the thickness of the steel layer and its permeability have a great effect upon the magnitude of Z_{12}.

Tables 7-1 and 7-2 give calculated values of Z_{12} for bimetallic shells of various structures.

№ обо-лочки (A)	Толщина медной оболочки, см (B)	Толщина стальной оболочки, см (C)	Магнитная проницаемость стали (D)
1	0,035	0,02	30
2	0,045	0,02	30
3	0,035	0,03	30
4	0,045	0,03	30
5	0,035	0,02	100
6	0,045	0,02	100
7	0,035	0,03	100
8	0,045	0,03	100
9	0,035	0,02	300
10	0,045	0,02	300
11	0,035	0,03	300
12	0,045	0,03	300

TABLE 7-1. Structural Data for Bimetallic Shells. A) Shell No.; B) thickness of copper shell, cm; C) thickness of steel shell, cm; D) permeability of the steel.

We have said above that the coupling impedance Z_{12} is numerically proportional to the current-density vector at the external surface of the outer conductor, $\dot{j}^c$. Similarly, it may be said that the current-density vector at the inner surface of the outer conductor is characterized by its proper impedance

$$Z_{в} = \rho \frac{j^{в}}{I}.$$

Thus the coupling impedance and the impedance of the outer conductor of the cable are related by the current densities or the longitudinal components of the electrical field at its external and internal surfaces:

$$\frac{Z_{12}}{Z_{б}} = \frac{j^{c}}{j^{б}} = \frac{E_z^{c}}{E_z^{б}} \qquad \text{or} \qquad Z_{12} = Z_{б}\frac{j^{c}}{j^{б}} = Z_{б}\frac{E_z^{c}}{E_z^{б}}.$$

The impedance of the outer conductor is defined by the formula

$$Z_{б} = \frac{\sigma}{\gamma_1}\cdot\frac{1}{2\pi b}\cot\sigma t. \qquad (7\text{-}3)$$

In the very high frequency region, where $|\sigma t| \geqslant 3$, the value of $|\coth \sigma t| \rightarrow 1$ and the formula is simplified to:

$$Z_{б} = \frac{\sigma}{\gamma_1}\cdot\frac{1}{2\pi b} \qquad (7\text{-}4)$$

It should be noted that the numerical value of Z_{12} does not change whether or not the source of energy lies within the cable (disturbing circuit) or outside it (disturbed circuit), while the value of Z_b does depend on this fact. From formula (7-4) Z_b may be calculated for the disturbing circuit. In order to calculate the proper impedance of the disturbed circuit, Z_c, the inside radius of

the outer conductor, b, should be replaced by its outer radius c in formula (7-4).

$$Z_c = \frac{c}{\gamma_1} \cdot \frac{1}{2\pi c} \cot \sigma t, \tag{7-5}$$

$$Z_c = \frac{c}{\gamma_1} \cdot \frac{1}{2\pi c}. \tag{7-6}$$

TABLE 7-2. Resistance of Various Types of Coaxial-Cable Sheaths, ohms/km. A) f, Mc; B) sheath number.

(B) № оболочки \ (A) f, Мгц	0,06	0,12
1	0,74	0,34
2	0,52	0,20
3	0,35	0,12
4	0,23	$0{,}7\cdot10^{-1}$
5	0,23	$0{,}6\cdot10^{-1}$
6	0,16	$4{,}1\cdot10^{-2}$
7	$0{,}5\cdot10^{-1}$	$1{,}1\cdot10^{-2}$
8	$0{,}4\cdot10^{-1}$	$0{,}7\cdot10^{-2}$
9	$3{,}5\cdot10^{-2}$	$4{,}6\cdot10^{-3}$
10	$2{,}4\cdot10^{-2}$	$2{,}8\cdot10^{-3}$
11	$3{,}6\cdot10^{-3}$	$1{,}9\cdot10^{-4}$
12	$2{,}4\cdot10^{-3}$	$1{,}1\cdot10^{-4}$

Table 7-2
Continued

0,25	0,5	1,0	2,0	4,0	8,0
$0,9\cdot10^{-1}$	$1,2\cdot10^{-3}$	$0,7\cdot10^{-3}$	$0,9\cdot10^{-5}$	$2,0\cdot10^{-8}$	$2,7\cdot10^{-12}$
$4,1\cdot10^{-2}$	$4,2\cdot10^{-3}$	$1,5\cdot10^{-4}$	$1,1\cdot10^{-6}$	$1,0\cdot10^{-9}$	$3,8\cdot10^{-14}$
$2,2\cdot10^{-2}$	$1,6\cdot10^{-3}$	$3,7\cdot10^{-5}$	$1,5\cdot10^{-7}$	$9,6\cdot10^{-10}$	$0,7\cdot10^{-15}$
$1,0\cdot10^{-2}$	$0,5\cdot10^{-3}$	$0,8\cdot10^{-5}$	$1,8\cdot10^{-8}$	$3,0\cdot10^{-11}$	$1,0\cdot10^{-17}$
$0,8\cdot10^{-2}$	$4,2\cdot10^{-4}$	$0,7\cdot10^{-5}$	$1,1\cdot10^{-8}$	$1,3\cdot10^{-12}$	$3,0\cdot10^{-18}$
$3,8\cdot10^{-3}$	$1,4\cdot10^{-4}$	$1,3\cdot10^{-6}$	$1,2\cdot10^{-8}$	$0,6\cdot10^{-13}$	$4,3\cdot10^{-20}$
$0,6\cdot10^{-3}$	$1,0\cdot10^{-5}$	$2,7\cdot10^{-8}$	$0,6\cdot10^{-11}$	$3,2\cdot10^{-17}$	$1,0\cdot10^{-24}$
$2,6\cdot10^{-4}$	$3,3\cdot10^{-6}$	$6,0\cdot10^{-9}$	$0,7\cdot10^{-12}$	$1,4\cdot10^{-15}$	$1,3\cdot10^{-26}$
$1,7\cdot10^{-4}$	$1,7\cdot10^{-6}$	$2,4\cdot10^{-9}$	$1,8\cdot10^{-13}$	$2,5\cdot10^{-19}$	$1,0\cdot10^{-27}$
$0,8\cdot10^{-4}$	$5,8\cdot10^{-7}$	$5,0\cdot10^{-10}$	$2,1\cdot10^{-14}$	$1,1\cdot10^{-20}$	$1,4\cdot10^{-29}$
$1,7\cdot10^{-6}$	$2,6\cdot10^{-9}$	$2,4\cdot10^{-13}$	$4,1\cdot10^{-19}$	$2,5\cdot10^{-27}$	$5,2\cdot10^{-29}$
$0,8\cdot10^{-6}$	$0,9\cdot10^{-9}$	$5,4\cdot10^{-14}$	$5,0\cdot10^{-20}$	$1,3\cdot10^{-28}$	$0,7\cdot10^{-40}$

7-3. CALCULATION OF THE CROSS-TALK ATTENUATION OF COAXIAL CABLES

The cross-talk attenuation between coaxial circuits depends upon the frequency of the current transmitted, the geometrical dimensions and material of the outer conductors, their location with respect to one another, and the impedance involved in completing a third intermediate circuit.

For the case in which the coaxial pairs are in direct contact (the most frequency encountered), the cross-talk attenuation per kilometer of cable is calculated according to formula (7-7); for short lengths of cable the cross-talk attenuation at the near end, B_0, and at the far end, B_l, are equal

$$B_0 = B_l = \ln\left|\frac{2ZZ_3}{Z_{12}^2}\right| = \ln|N|, \qquad (7\text{-}7)$$

where Z is the wave impedance of the coaxial cable; Z_{12} is the coupling impedance in ohms/km; Z_3 is the impedance of the intermediate circuit formed by the two external shells of the coaxial pair under consideration, in ohms/km.

For long cable lines, more than 1 km long, the cross-talk attenuation may be calculated on the basis of the following formulas:

a) for a comparatively short section of cable line (where $\beta l \ll 1$) the cross-talk attenuation at the near end, B_{0n}, and at the far end, B_{ln}, of the cable is the same and equals:

$$B_{0n} = B_{ln} = \ln\left|\frac{2ZZ_9}{Z_{12}^2 l}\right| = \ln\left|\frac{N}{l}\right|, \qquad (7\text{-}8)$$

where l is the length of the cable, in km;

b) for long cable lines (where $\beta l > 1$) the cross-talk attenuation at the near end is:

$$B_{0n} = \ln\left|\frac{2ZZ_9}{Z_{12}^2} 2\gamma\right| = \ln|N \cdot 2\gamma|, \qquad (7\text{-}9)$$

where γ is the propagation constant of the coaxial cable.

The resistance to interference between the coaxial circuits is

$$B_{12} = \ln\left|\frac{2ZZ_9}{Z_{12}^2 l}\right| = \ln\left|\frac{N}{l}\right|. \qquad (7\text{-}10)$$

The cross-talk attenuation at the far end is:

$$B_{ln} = B_{12} + \beta l = \ln\left|\frac{2ZZ_9}{Z_{12}^2 l}\right| + \beta l = \ln\left|\frac{N}{l}\right| + \beta l, \qquad (7\text{-}11)$$

where β is the attenuation constant of the coaxial cable.

Comparing the formulas for computing the cross-talk attenuation and interference resistance of the symmetric and coaxial cables, we may note that they are identical in form. The difference lies in the fact that, in the symmetrical circuits, owing to the indefiniteness of the phase shifts introduced by the individual sections of the cable, the interaction adds geometrically, and the length of the line enters into the expressions for B_{1n} and B_{12} under a radical ($\sqrt{n}$ or $\sqrt{l}$), while in the joined sections of coaxial circuits the phases have the same sign which permits arithmetic addition of the interaction, and the length l enters into the formulas directly.

The quantities Z, γ, β in expressions (7-7)-(7-11) are calculated on the basis of the formulas presented in section 5-5. The coupling impedance Z_{12} is determined in the manner shown in the section 7-1.

The impedance of the intermediate circuit, Z_3, consists of the impedance Z_c of the outer conductors of both coaxial cables, and the inductive reactance of the circuit formed by them:

$$Z_3 = 2Z_c + j\omega L_3, \tag{7-12}$$

where L_3 is the external inductance, equal to:

$$L_3 = 4 \ln \frac{2a' - D'}{D} \cdot 10^{-4} \text{ [Henrys/KM]} \qquad (7\text{-}13)$$

All the symbols given in (7-13) are shown on Fig. 7-10.

In the case where the coaxial pairs are in contact, the external inductance $L_3 = 0$, and the impedance of the intermediate circuit equals the sum of the impedances of the outer conductors of the cables.

$$Z_3 = 2Z_c. \qquad (7\text{-}14)$$

The quantity Z_c includes both the pure resistance and the reactance introduced by the internal inductance of the outer conductor of the cable. The formulas for computing Z_c are given in section 7-2.

Tables 7-3 and 7-4 give the results of measurements of cross-talk attenuation at 60 kc between pairs of cables. The cable used was type 2.6/9.4, 425 m long, consisting of four coaxial circuits, type 1.83/6.7, 16 km long, and type 3.17/11.7, 12 km long.

Fig. 7-11 gives the frequency dependence of the cross-talk attenuation at the near and far ends for type 5/18 coaxial cables.

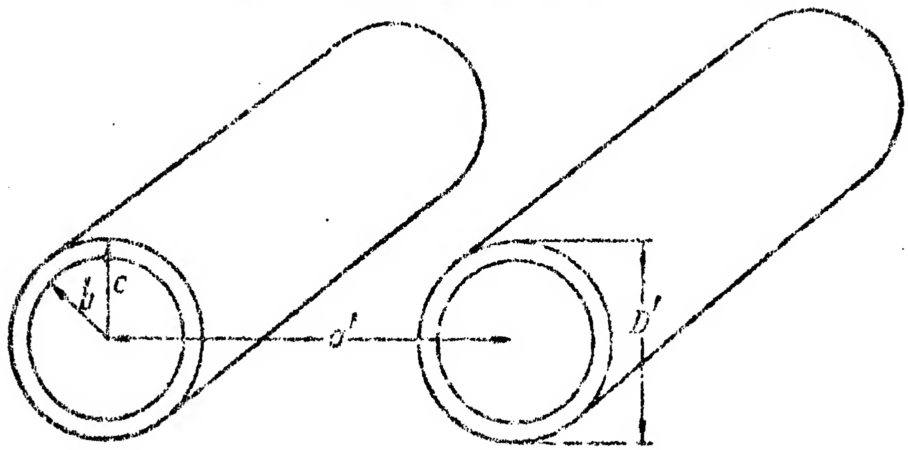


Fig. 7-10. Diagram for the computation of cross-talk attenuation between coaxial pairs.

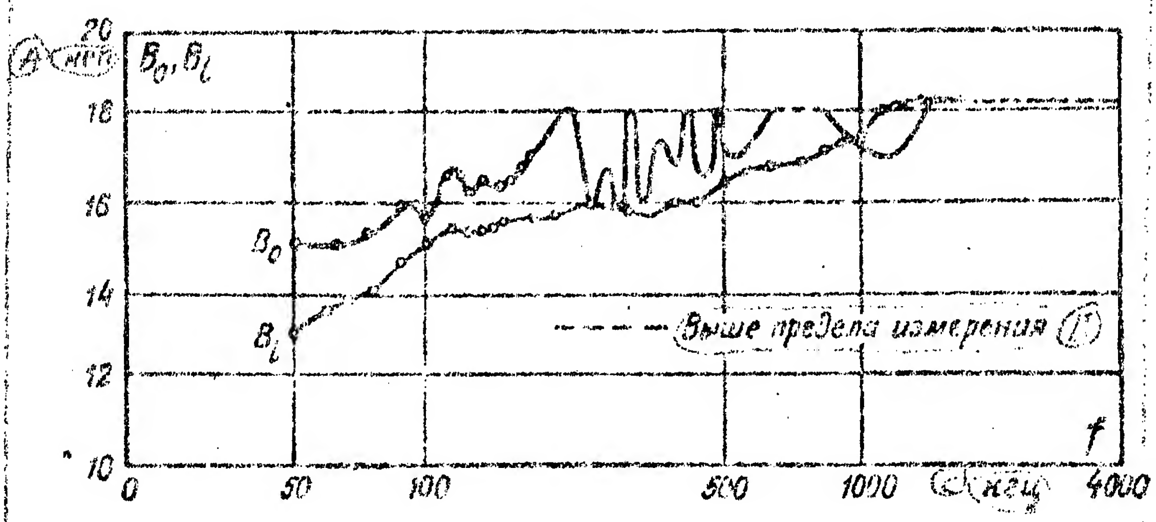


Fig. 7-11. Frequency dependence of cross-talk attenuation at the near end, B_0, and far end, B_1, for type 5/18 coaxial cables. A) Nepers; B) above the limit of measurement; C) kc.

It follows from the data that has been presented that the cross-talk attenuation in coaxial cables is greater at the near end than at the far, in contrast to the situation existing with symmetric construction. The cross-talk attenuation increases with frequency, and the interference resistance of the coaxial cables increases too.

TABLE 7-3.

A Между какими парами	Переходное затухание на ближнем конце, B неп	Переходное затухание на дальнем конце, C неп
I/II	12,0	9,7
I/III	12,2	10,9
I/IV	11,8	10,8
II/III	11,7	10,7
II/IV	12,0	10,8
III/IV	11,7	11,7

Results of Measuring Cross-Talk Attenuation Between Coaxial Pairs of a Type 2.6/9.4 Cable (f = 60,000 cps). A) Pairs across which measurements were made; B) cross-talk attenuation at the near end, nepers; C) cross-talk attenuation at the far end, nepers.

TABLE 7-4.

Частота, кгц (A)	Кабель 1,83/6,7 Длина 16 км (B)		Кабель 3,17/11,7 Длина 12 км (C)	
	B_0, неп (D)	B_1, неп (E)	B_0, неп (F)	B_1, неп (G)
50	14,5	11,5	—	—
100	15,85	13,2	12,7	10,5
200	18,44	14,95	14,5	12,7
300	—	—	15,7	14,3
400	—	—	—	15,3
500	—	—	—	16,0

Results of Measurement of Cross-Talk Attenuation in Long Coaxial Lines. A) Frequency, cps; B) Type 1.83/6.7 cable, 16 km long; C) type 3.17/11.7 cable, 12 km long; D) B_0, nepers; E) B_1, nepers; F) B_0, nepers; G) B_1, nepers.

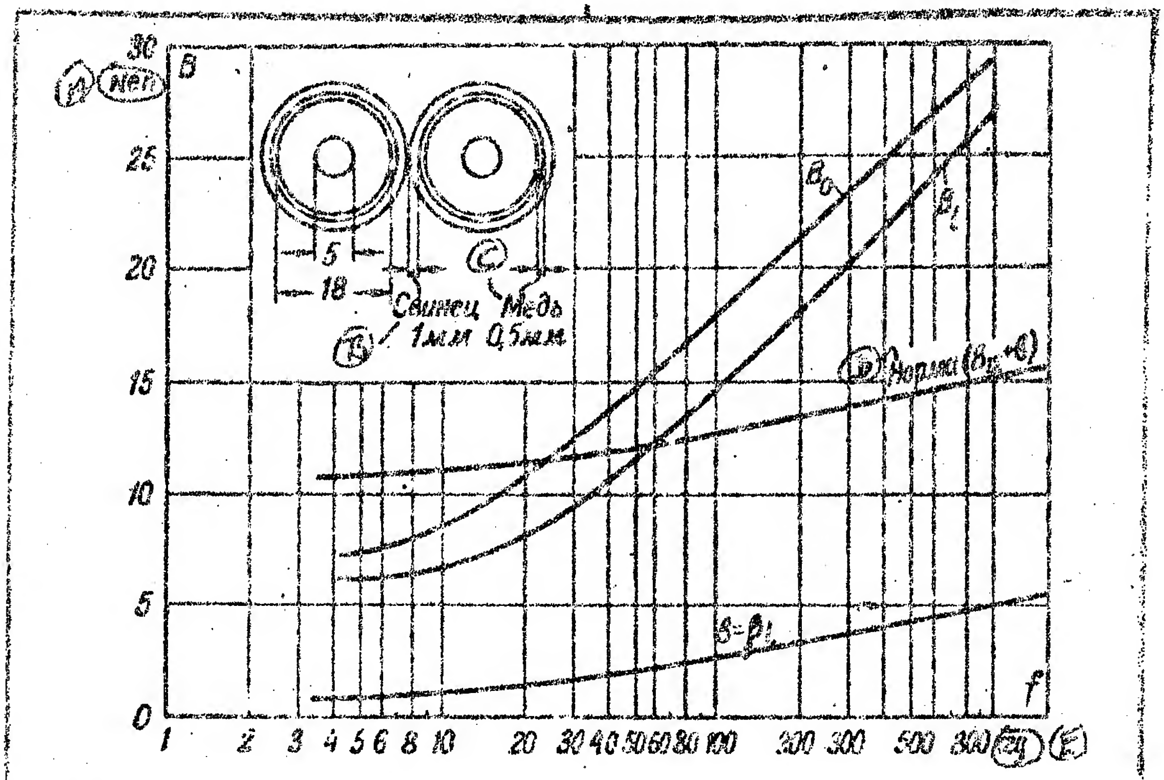


Fig. 7-12. Frequency dependence of cross-talk attenuation between two 35-km long coaxial cables. A) Nepers; B) lead; C) copper; D) standard; E) cps.

Fig. 7-12 gives the values of the cross-talk attenuation B_0 and B_1 between two type 5/18 coaxial cables, equal in length to a repeater section — 35 km. The same figure gives the standard curve which must be satisfied by a coaxial cable in accordance with the recommendation of the International Consultative Committee. According to the existing norms of the committee the ability of coaxial circuits to reject noise must be not less than 9.8 nepers at all frequencies used. Allowing for the attenuation of the

cable itself the standard cross-talk attenuation will be $B_{12} + \beta l$.

It is clear from Fig. 7-12 that a coaxial cable begins to satisfy the cross-talk attenuation requirements only at frequencies of 60 kc and above.

From Fig. 7-13, where the dependence of the values of B_{0n}, B_{ln}, and B_{12} upon line length is given, it is clear that as the length increases the ability of coaxial circuits to reject noise decreases (B_{12} decreases). The cross-

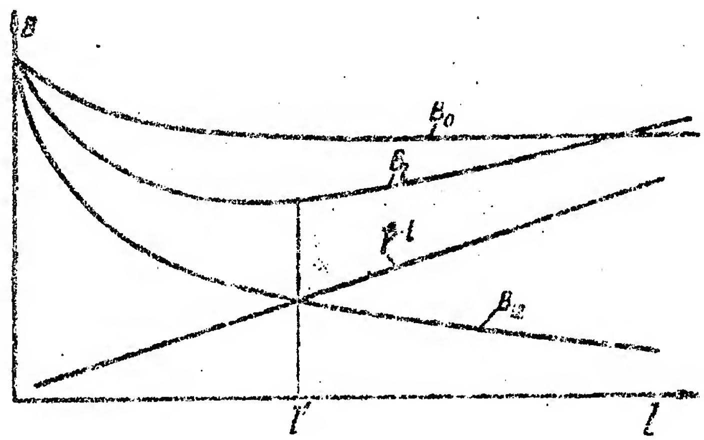


Fig. 7-13. Variation in the values of cross-talk attenuation with the length of a coaxial cable line.

talk attenuation at the near end, which drops somewhat at first, is constant, and equals $B_{0n} = \ln |N \cdot 2\gamma|$. The cross-talk attenuation at the far end has a minimum at a

specific distance (l'), and then rises as the length of the line increases; the reason for this is the increase in the attenuation l of the cable itself.

7-4. INTERACTION BETWEEN COAXIAL PAIRS LOCATED WITHIN A COMMON LEAD SHEATH

The computational formulas which have been given above allow only for direct interaction between two isolated coaxial pairs. In actual cables, in addition to coaxial pairs, there are third circuits. We refer to the lead sheathing and armoring within which the coaxial pairs are located, as well as the remaining circuits contained within the single cable.

In general, the lead sheathing and third circuits act to favor coaxial cables with respect to interaction, increasing the cross-talk attenuation between the coaxial circuits and improving their ability to reject noise.

This is explained physically by the shielding action of the third circuit. They carry off some of the interaction energy and increase the noise resistance of the disturbed circuit. Here the effect is completely similar to that in the case of a shielding wire used to decrease the interference of aerial power-transmission lines with a communication line.

As can be seen from Fig. 7-14, the interfering circuit I induces emfs and corresponding noise currents I_2' and I_3' in the disturbed circuit II and in the shield wire.

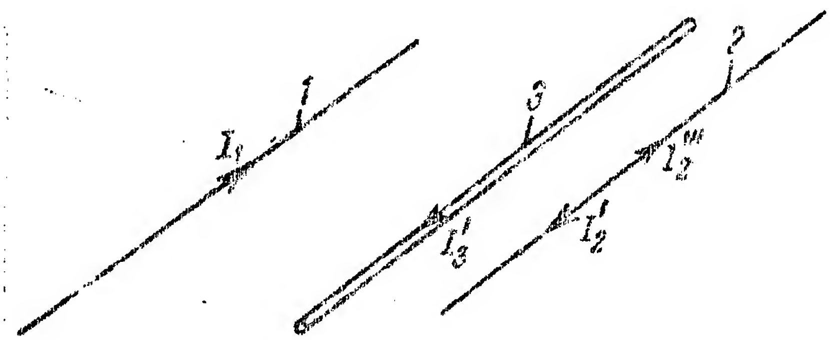


Fig. 7-14. Principle of a shield wire. 1) Disturbing circuit; 3) shield wire; 2) disturbed circuit.

The current I_3' set up in the shield wire in turn induces an emf, and sets up a current I_2'' in circuit II. This current is opposite in direction to the current I_2'.

As a result, the difference current $I_2' - I_2''$ acts in circuit II, and the magnitude of the interference is less in the presence of the shield wire than in its absence.

In actuality the mechanism of interaction in coaxial cables in the presence of third circuits is quite complicated; however, the example given is adequate to show what is happening physically in the screening action of third circuits. In a combination coaxial cable, the lead

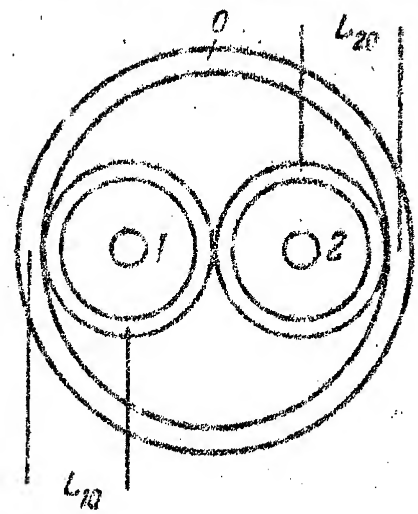


Fig. 7-15. Calculating the interaction between coaxial circuits in the presence of a common lead sheath. 1) First coaxial circuit; 2) second coaxial circuit; 0) lead sheath.

sheathing plays the role of a shield wire, as do the armoring and other third circuits included in the cable.

In the general case, the calculation of interaction in combined cables is rather complicated. For practical purposes, the greatest interest is presented by the case of the interaction of two coaxial pairs located symmetrically within a common lead sheath (Fig. 7-15).

In this case the cross-talk attenuation may be expressed in the terms of the cross-talk attenuation between the two separate coaxial circuits, B, and the additional

cross-talk attenuation, due to the lead sheathing, acting as a third circuit, B_3

$$\text{overall } B_{\text{общ}} = B + B_3. \qquad (7\text{-}15)$$

The way in which B is determined has been explained above.

The value of B_3 may be computed approximately by using the following formula:

$$B_3 = \ln \left| \omega \frac{L_{10} L_{20}}{L_{10/20}} \right|, \qquad (7\text{-}16)$$

where L_{10} is the inductance of the circuit: coaxial pair I — lead sheath;

L_{20} is the inductance of the circuit: coaxial pair II — lead sheath;

$L_{10/20}$ is the mutual inductance of the circuit: coaxial pair I — sheath and pair II — sheath.

Assuming that these parameters have approximately the following values: $L_{10} = L_{20} \approx 2 \cdot 10^{-3}$ henrys/km and $L_{10/20} = 4 \cdot 10^{-6}$ henrys/km, the additional cross-talk attenuation may be computed from the formula:

$$B_3 \approx \ln \omega. \qquad (7\text{-}17)$$

Consequently, the higher the frequency, the higher the cross-talk attenuation introduced by the common lead sheath, and the higher the "shielding" effect of third circuits.

2184

END